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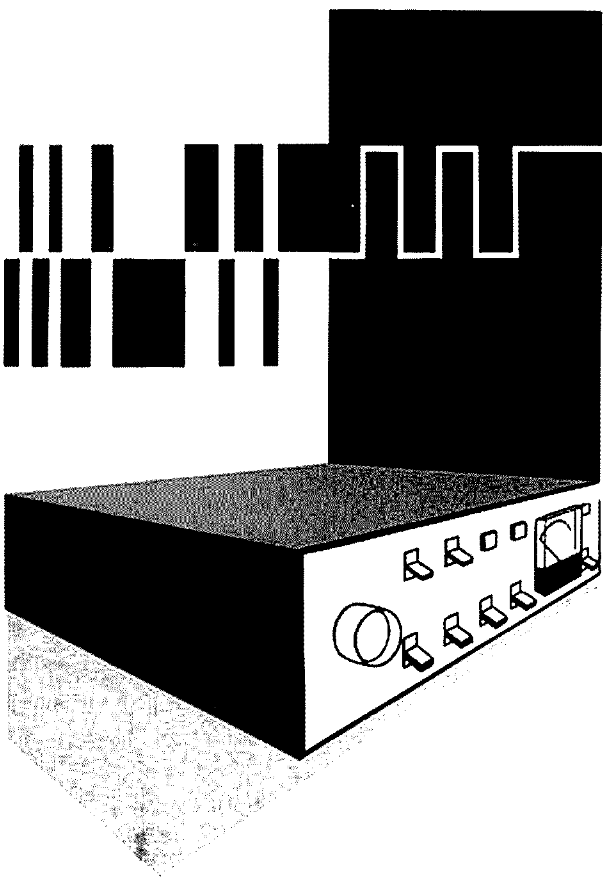
ham radio magazine

JANUARY 1971

THE MAINLINE ST-6 RTTY DEMODULATOR



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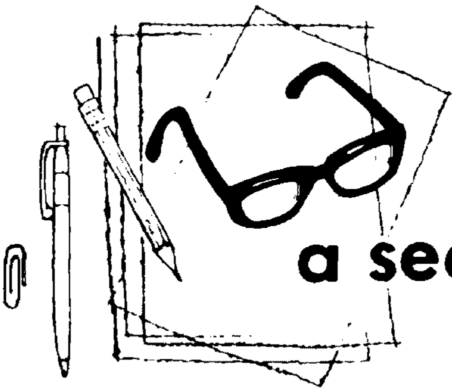


a second look

by jim fisk

Although the number of manned Apollo missions to the moon has been drastically cut back by NASA, the amateur group working on Moonray still hopes to have their uhf lunar repeater carried to the moon on one of the remaining flights. The primary objective of Moonray, proposed in 1960 by W6OLO, is to provide a free-access uhf repeater for world-wide communication on the 432-MHz band.

The Moonray package will contain a sensitive low-noise receiver, tuned to 439.9 MHz, a signal processor, an identifier, six to eight channels of telemetry, and transmitter output on 430.1 MHz, as well as a laser receiver. Power will be provided by a nuclear thermoelectric generator with a half life of 87 years. The system is expected to be continuously operational for a year or longer on a 24-hour-per-day basis, with one-minute interruptions at the end of each 10-minute period.

The antenna, pointing system and thermal controls are all self contained in a louvered metal cylinder about 6 inches in diameter and 10 inches long. Three legs will be used for supporting the package on the lunar surface, as well as leveling; a special pointing mechanism is provided so the astronaut who sets up the unit will be able to accurately align the Moonray antenna to earth.

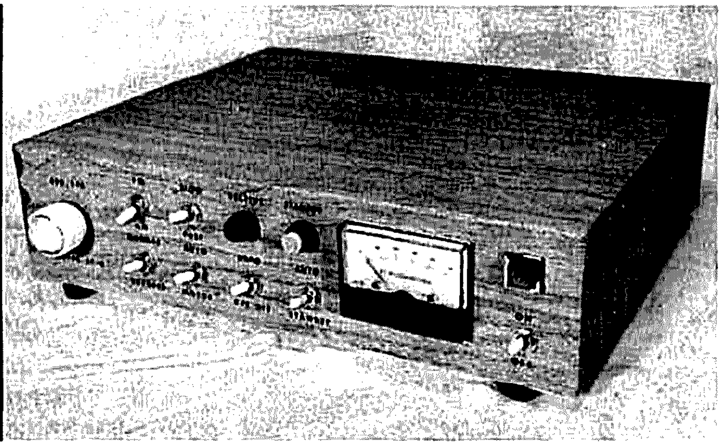
The 10-KHz passband on both the 430.1- and 439.9- MHz links will accept all modes of transmission, including cw, fsk, modulated cw, afsk, narrow-band fm, ssb and a-m (in that order of preference). Moonray's callsign will be the identifier SS, transmitted every ten minutes along with a telemetry sequence.

Amateurs who want to use the Moonray repeater should be equipped with a high-gain antenna (15-dB minimum gain) capable of tracking the moon, as well as a low-noise (3-dB maximum) crystal-controlled converter on 430.1 MHz, and a stable 439.9-MHz transmitter. For cw operation, transmitter power output should be approximately 50 watts. More power will be required for other modes, progressing from about 100 watts for fsk to 1000 watts for slow-scan television.

The design, development and test of an operational prototype is now in progress, and final construction of a flight model is planned for the near future. There are many ways you can help with this ambitious project including technical help to the group, providing publicity, promoting additional 432-MHz activity, or by making financial contributions to help obtain special items and defray postage and printing costs (contributions are fully tax deductible).

To obtain more technical information on Moonray, or to offer help, write to Nassau College Amateur Satellite Tracking, Astronomy and Radio, Post Office Box T, Syosset, New York 11791. Your help will be appreciated.

Jim Fisk, W1DTY
editor



the Mainline ST-6 RTTY demodulator

Here's an ultra-modern RTTY tuning unit that features the latest in electronic circuitry and design

Irvin M. Hoff, W6FFC, Los Altos Hills, California 94022

An RTTY demodulator (often called a terminal unit, TU or converter) is used between the radio receiver and the Teletype machine to change the audio output of the receiver into on-off dc pulses to operate the printer.

Most RTTY demodulators work on the fm principle – they have a limiter followed by a discriminator that selects the mark or space tone, a detector to convert this audio into the dc keying rate, perhaps a low-pass filter to clean up the signal, a trigger stage that decides when mark has changed to space and vice versa, and finally, a keyer stage that drives the Teletype machine. Two power supplies are normally used, one to drive the demodulator and one to drive the Teletype machine, called the loop supply.

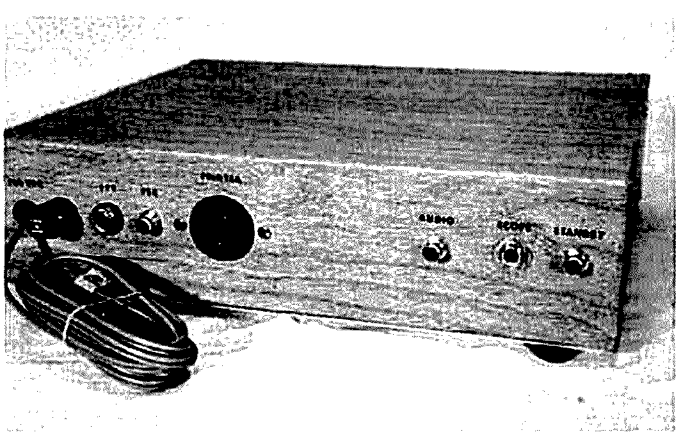
Some demodulators work on the a-m principle. This is often called two-tone, or limiterless operation. In this case, no limiter is used although some modest amplification might be incorporated. A threshold corrector is required to provide

the proper symmetry to the slicer. This type of demodulator can copy mark-only or space-only type signals also. It works quite well in some circumstances, especially at full automatic machine speed, sometimes called "tape speed." To realize the advantages of a-m reception, high-performance channel filters should be used, but rarely are — a simple linear discriminator providing little advantage over fm in a majority of cases. If automatic printer control (auto-start) has been added, it must be disabled for a-m copy since the steady reference level used for the squelch system is no longer a stable voltage.

A good demodulator should have provisions for both fm and a-m operation, as the circuitry used for a-m is beneficial even when using normal fm reception.

the mainline rtty demodulators

The original Mainline RTTY converter was published in 1963.[1] The Mainline TT/L was introduced in late 1964[2] and had features that soon set the standard for serious RTTY enthusiasts. The Mainline TT/L-2[3] was an up-graded TT/L with a heftier power and loop supply, plus a few minor circuit improvements.



Rear-panel layout of the ST-6 RTTY demodulator.

Then I turned to solid-state — all units just mentioned used vacuum tubes. The first in the new series was the ST-1, using an RCA integrated circuit. It was never published, as the GE PA-238 operational amplifier became available about the same time. The ST-2 (autostart but no motor delay) and the ST-3 (autostart with motor delay) were published in 1968.[4] The ST-4 was the same as the ST-3, but designed exclusively for narrow-shift (170 Hz) reception for autostart. The ST-5[5] used two 709C operational amplifiers and was as simple a unit as a person could hope to get for normal reception of RTTY signals. The Mainline ST-6 presented here is an advanced solid-state demodulator offering all the practical advantages known to be beneficial for RTTY reception.

features of the ST-6

1. Optional bandpass input filters of 3-pole Butterworth design.
2. Use of seven 709C operational amplifiers; nine if both 170 and 850 shifts are added.
3. A limiter having over 90-dB gain.
4. Well-designed linear discriminators for various shifts.
5. Full-wave detection.
6. Active low-pass minimum bandwidth 3-pole Butterworth filter using two operational amplifiers.
7. Threshold corrector for a-m reception.
8. High-gain operational amplifier for a slicer that allows reception on shifts as low as 1 Hz.
9. 300-volt keying transistor stage for minimum distortion.
10. Floating-loop supply giving optimum plus-minus voltages for best FSK (transmitter) provisions.
11. Narrow-shift cw identification provisions.
12. Regulated plus-minus 12-volt supply.
13. *Anti-space* system that quickly puts the printer to mark-hold if the signal goes to steady space.
14. Automatic printer control (auto-start) that ignores voice or cw signals but responds to RTTY, turning on the printer motor automatically.
15. Motor delay control to keep the motor from turning on and off exces-

sively during weak signal reception or compulsory cw identification.

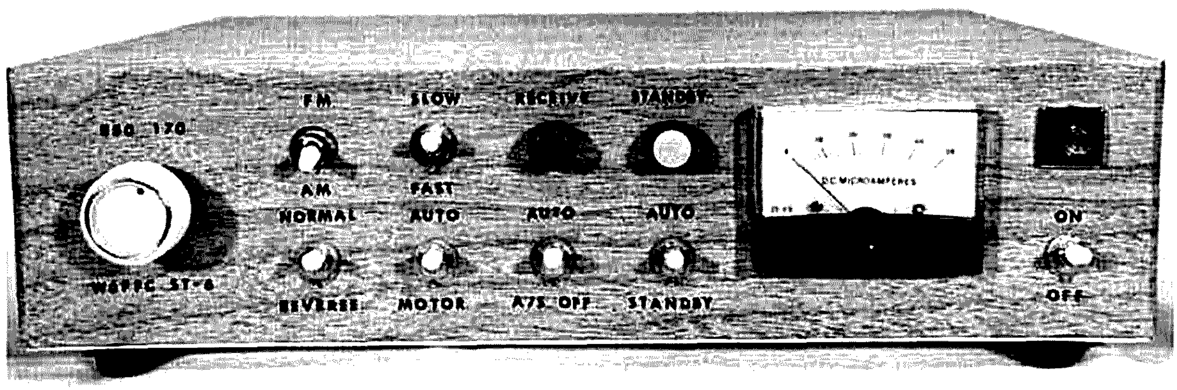
16. Optional a-m or fm reception.
17. Simplified switching for versatility.
18. Low-cost but effective tuning indicator.
19. Very simple alignment and adjustment procedures.

circuit description

The bandpass input filters These are 3-pole Butterworth design, based on modern filter synthesis networks. The one for 850 shift uses normal 88-mH toroids wired in series for 88 mH. It is about 1 kHz wide at the ±3 dB points. The filter for 170 shift uses 88-mH toroids wired in parallel for 22 mH. (The two outer wires together and the two pigtails together.) This keeps the impedance of the two units similar. The 170-shift input filter is about 275 Hz wide.

The limiter This is a 709C operational amplifier running "open loop" for maximum gain. It has been frequency compensated only lightly. For a-m reception the gain is reduced so that the unit becomes a linear amplifier at normal input levels, but reverts to a controlled limiter if the input exceeds the normal level.

The discriminator When using inductors of the same value for both mark and space, great attention must be given by the designer in order to achieve similar bandwidths, similar voltages, good zero-axis crossover, good noise balance, etc. It becomes even more difficult to design discriminators for both 170 and 850 which can be interchanged in the same overall circuit. Attention to these details has been given to all the Mainline demodulators and the ST-6 is based on the experience learned from the other units.



Front-panel layout of the ST-6 RTTY demodulator.

The detector Full-wave detection is used, a unique feature of the Mainline ST-series demodulators. This offers optimum detection for easiest filtering of the remaining ripple component. An additional "plus-plus" detector is incorporated for a tuning display and to drive the automatic printer control system.

The lowpass filter The optimum lowpass filter of minimum bandwidth will do more to improve the performance of a demodulator than any other single thing, assuming a high-voltage loop supply is already being used. Vic Poor[6] pointed out in his outstanding article on RTTY filters that this filter should cut off at

27.3 Hz for 60 speed (45.45) Baud operation, and set up the criteria for testing such filters. The active lowpass filter in the TT/L was built with these features in mind. Fig. 1 shows the curve obtained with the ST-6. The observed "eye pattern" was ideal, indicating optimum cutoff for 60 wpm. The active filter is one that uses feedback circuits with amplification rather than passive components (inductors, etc.). This gives consistent uniformity from one unit to

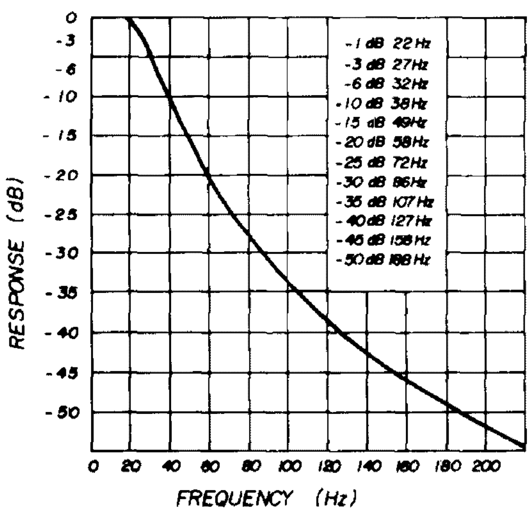


fig. 1. Bandpass curve for the low-pass active filter used in the ST-6.

another, small size, modern design, light weight, low-cost components and freedom from magnetic fields and hum.

The atc threshold corrector This section provides the automatic symmetry needed for the slicer to minimize distortion when using limiterless a-m reception. It also enables mark-only or space-only reception. It is beneficial even with the limiter in use as it offers a form of diversity reception, capitalizing on the redundancy of the mark and space signals. As a good minimum-bandwidth lowpass filter changes square-wave inputs to sine-wave outputs, such a threshold corrector is practically a requirement for proper zero-axis crossover at the slicer.

The slicer This is another 709C operational amplifier running in "open loop" for maximum gain. It is heavily frequency compensated since the signals handled are at nearly dc. The output of the lowpass filter is so clean and the gain of this stage so high that signals as low as 1 Hz shift will adequately swing the output from full mark to full space. No hysteresis was either necessary or desired.

The control section The output of the slicer is followed by a simple control section that allows a mark-hold voltage to be placed on the keyer stage when there is no authentic RTTY signal, putting it in mark-hold. This same control section allows simple use of the Sel-Cal[7] unit some enthusiasts have built for autostart purposes. It responds only to their own call letters. The control section is part of the manual standby system.

The loop supply This is a 180-volt supply that gives virtually distortion-free keying of the Teleprinter. A voltage-absorbing network is used on the collector of the keying transistor to protect it against excessive back-emf of the printer magnets which could otherwise occur if two or three printers were used on the same loop supply. The loop also puts out plus-minus voltages at the FSK output for best transmitter keying. It also adapts instantly (with no changes) to the Mainline AK-1 AFSK Keyer[8] for ssb units. It also offers narrow-shift cw identification.

Normal loop current is approximately 60 mA. This means the selector magnets of the printer and/or reperf should be in parallel. This keeps the inductance in the circuit at 25% the value it would be were the selector magnets in series. If a reperf and printer are to be in the same circuit at the same time, they would then go in series with each other, both having their own selector magnets in parallel. For solid-state keyers the lower the inductance the better due to "spikes" generated by the back-emf of the inductors in a collapsing field.

The power supply This plus-minus 12-volt supply is transistor and zener-regulated for good stability. It is heavily fused to

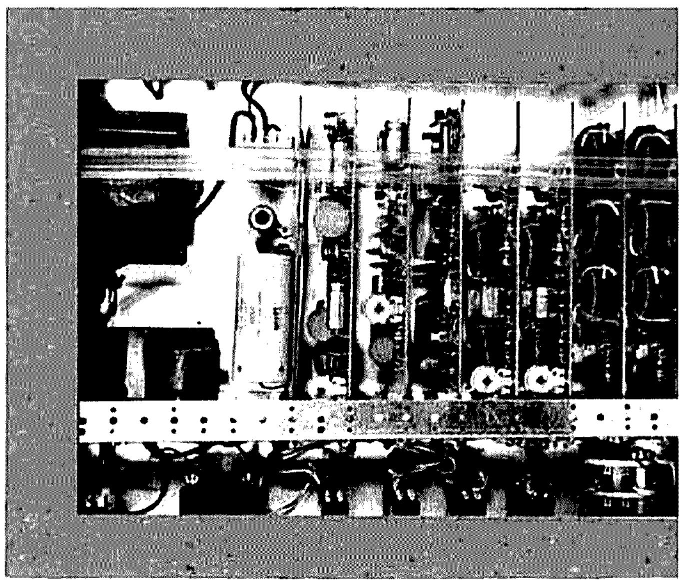
prevent any damage during long periods of unattended operation and is bypassed for rf.

The anti-space circuit A "blank" key has the longest continuous amount of space time of any RTTY character, about 132 milli-seconds. Any space signal longer than this cannot be RTTY. The anti-space system samples the mark-space output of the slicer. If the signal goes to space more than 132 ms the system says "tilt" and locks the printer to mark-hold, at the same time turning the automatic printer control off and starting the countdown on motor delay cutoff. Thus, if a person is checking his shift at the other end, your printer does not sit there running "open." This also gives greatly improved protection from a non-RTTY signal in the space channel from activating the automatic printer control and turning the motor on. This system works just as effectively with the limiter on or off, with the autostart on or off and with straddle tuning of any shift. No provisions for disabling it were provided. If this were needed, a switch to short the base of Q7 to ground would be suitable.

Tuning meter Although labeled as optional this is a most worthwhile device. Many amateurs prefer a scope for tuning, but this type of meter gives satisfactory results. It is better than most scope displays for straddle tuning where the incoming signal is not very accurately adjusted for correct shift, i.e. the operator might be using 600 shift, thinking it was in reality 850. The tuning meter is also helpful in adjusting the ST-6, although any voltmeter can be used for alignment.

Automatic printer control This system is normally called autostart. It is a unique squelch system that responds only to authentic RTTY, or to a steady mark carrier. A steady space carrier will not affect it due to the anti-space system. It samples both the mark and space channels for a signal having a duty time in excess of about 75%. Since most cw has a duty time of less than 50%, and most voice signals a duty time of perhaps under 30%, this system responds almost exclusively to RTTY signals.

To get the 75% duty time requirement an analog charging circuit is used. Since this is based on fm squelch principles, the use of limiterless a-m is not compatible with this system and is disabled if the limiter is turned off. Since momentary static crashes, deep fades and other temporary interference could cause the system to decide there was no longer any RTTY present, a one-second delay is incorporated into the shut-off feature. This in turn requires about a 3 to 4 second turn-on delay to obtain 75% recognition. Other ratios are shown in table 1 for those who may wish to experiment.



Internal construction of the ST-6 built by the author. A total of eight printed-circuit boards are used.

The higher ratios (like 8:1) will only respond to the very best RTTY signals and will not try to copy those signals that would normally produce garble, but would lock up if the signal deteriorated momentarily. Perhaps their biggest disadvantage is the long turn-on time: 6 to 7 seconds. This is the equivalent of about a half-line of RTTY.

The system used for automatic printer

table 1. Autostart ratios which may be used for protection from printing garble from marginal signals.

ratio	duty time	R61	R59	R60	time to on	time to off
2:1	67%	5.1k	390	4.7k	1.80	.88
3:1	75%	3.6k	2.4k	4.7k	2.48	.84
3.3:1	77%	3.6k	2.7k	5.1k	2.74	.86
4:1	80%	3.3k	3.9k	6.8k	3.53	.87
5:1	83%	3.0k	5.1k	6.8k	4.16	.84
6:1	86%	3.0k	6.8k	8.2k	5.25	.88
7:1	88%	3.0k	9.1k	9.1k	6.38	.90
8:1	89%	3.0k	10k	11k	7.35	.93

control in the Mainline series is unique in that it requires no special call-up or turn-off characters. As a result, it is extremely useful for normal RTTY QSOs where autostart itself would not be considered by the operator. In this case automatic printer control is a more vivid and realistic description of the system than autostart.

Motor delay This feature allows the automatic printer control to place the printer in mark-hold without turning the motor off. During poor conditions, it is a nuisance to have the motor turning on and off. During fast break operation the motor could turn off if it took more than approximately one second for the other station to respond. Also, the FCC requires cw indentification between transmissions, and this would normally cause the motor to turn off if no delay system were used. The motor delay in the ST-6 keeps the motor running for 20 to 30 seconds after the autostart system thinks there is no further RTTY present.

Loop supply cutoff A feature is incorporated in the ST-6 that opens the loop supply to space whenever the motor relay turns the motor off. Otherwise there is some 10 watts of heat in the cabinet from the 2750-ohm, 20-watt loop-current regulating resistor. This is not detrimental, but serves no useful purpose, and consumes an additional 10 watts of unnecessary power when the printer is not running.

Fuses The ST-6 probably has more fuses in it than any other item in your entire station. They aren't all *necessary,* but for unattended operation good fusing is worthwhile. Receivers, transmitters and other electronic items have been known to develop partial shorts in the high-voltage secondary systems, destroy components which may occasionally catch on fire, ruin transformers and *still not blow the fuse* in the primary! This is often because the secondary has several windings. By merely fusing the primary, it is hard to select a fuse that will quickly blow if the secondary pulls excessive current in *one* of the windings.

Thus the ST-6 fuses each secondary winding as well as the combined primaries. The jack for the motor is not fused (and should not go through the on-off switch) since the printers have their own protective fuses. The relay handles the motor current.

On the printed-circuit boards from Halle,* the indicator lamps also run through the fuses. In the event you add those lamps, you may find it necessary to increase the size of the fuses somewhat.

Rf protection The operational amplifiers have such fantastic gain that precautions should be taken to prevent amplification of nearby rf signals, because their usefulness is destroyed while that rf is present. The ST-6 bypasses the 120 Vac input line, the power supply output, and isolates the plus-minus voltages to each op amp to prevent any instability or oscilla-

tion. Thus, no adverse affects have been noted from nearby kilowatt transmitters.

Scope connections Provisions have been included for connecting an oscilloscope. the 1 megohm isolating resistors provide adequate protection against the low-impedance scope inputs from affecting discriminator balance. If you incorporate a scope it should be an external unit that plugs into the rear of the ST-6. This allows the scope to be used for other demodulators, to be larger than the 1" or 2" display the ST-6 cabinet might allow, and allows the operator to run the ST-6 twenty-four hours a day for autostart.

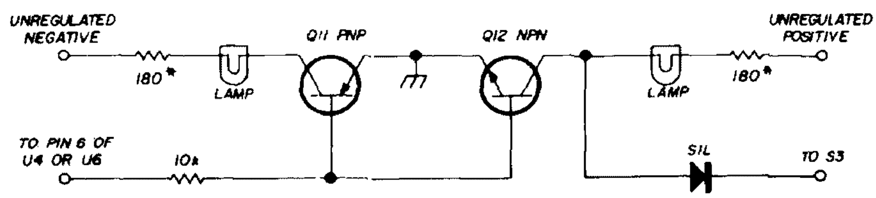


fig. 3. Simple lamp circuit to indicate standby and receive. Resistors marked with asterisk are selected for normal lamp current.

the switches

S1 is a dual-purpose switch to change the unit from fm to a-m reception. When in a-m limiterless, it also disables the autostart system, putting the motor automatically to *on.*

S2 is the normal-reverse switch. It is not necessary, but there are rare occasions when a newcomer is "upside-down" and you may want to see who it is. Many people will leave this switch completely off their unit.

S3 is the local standby switch. This places the printer in mark-hold and disables the autostart (turns the motor on and keeps it on). There is also a jack for an external remote standby switch. **Fig. 2** shows how this can be used. Switches S10 and S11 would be placed at or on the printer itself; S10 as a standby switch and S11 as the transmit control switch. This not only automatically turns the transmitter on, it mutes the ST-6 so incoming signals for the tuning meter or scope do not affect the transmitted signal – without this provision acoustical feedback could cause incorrect signals to be transmitted. By placing these two switches next to the printer single-switch station control may be used, and the ST-6 can be across the room out of arm's reach. This offers maximum versatility plus simple fast-break operation.

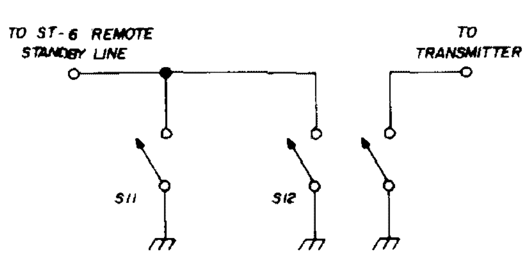


fig. 2. Station switching. S11 and S12 normally located near printer. Standby switch S11 keeps machine from printing; station control switch S12 puts the transmitter on the air and mutes the ST-6 for local copy.

S4 is the fast-slow autostart; it also turns the motor on when in *fast* and keeps it on. The *fast* system uses the same ratio of turn-on to turn-off times, but speeds them both up. This makes it convenient to use automatic printer control for fast break, but gives inadequate protection against momentary signal loss during unattended operation.

S5 merely turns the automatic printer control off, allowing manual operation via the standby switch S3 or S10. (S10 is not shown on the schematic, it is shown on **fig. 2.**)

S6 is the manual motor *on* switch. It keeps the motor running regardless of any other switch position except the power off switch, S7.

S7 turns the 120 Vac power on and off, and opens the FSK line when in off, to prevent any hum loops from the FSK keyer in the transmitter. This could be annoying when using voice or normal cw if the ST-6 were in the off position. This is normally no problem, but the switch is

available for this purpose if you want to use it.

standby-receive lamps

Fig. 3. shows how lamps may be added to indicate standby and receive. When the manual or remote standby switch is activated the standby lamp turns on, whether the receive lamp is on or not. The receive lamp works only from the output of the automatic printer control system.

Thus it is possible to have both lamps on at the same time. This tells the operator he has selected manual standby. As a result he is informed that he is not in automatic reception mode at all, and before leaving the room he must turn off the standby switch (S3 or S10) if he expects to get automatic copy. It makes an excellent fail-safe reminder.

If you prefer an indicator for mark and space signals, attach the 10k resistor in **fig. 3** to pin 6 of U4. In this case the diode on the collector of Q12 is removed.

850-170 switch

If incorporating both 850 and 170 shifts, duplicate the front end up to but not including the first 16k resistor on U3. Then a multipole two-position switch would be used to switch the audio input, the output of the two U2 stages, the output of the U1 limiter for the limiterless a-m mode, and the two 47k resistors on the inputs to the U1 stages. The scope display for mark and space probably should be switched as well as the output to the meter and autostart at point 2. Other poles can be used to ground the audio input of the unused limiter off. I used a two-section (6 poles per section) two-position ceramic switch.

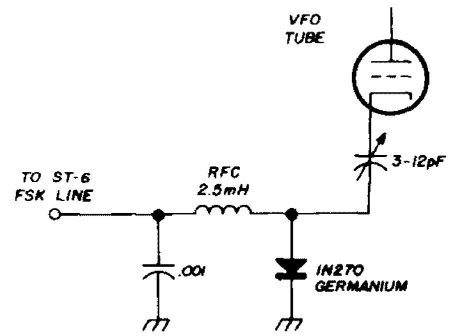


fig. 4. Circuit for adding frequency-shift keying to your transmitter. If your signal is "upside down," turn the diode around.

construction

The unit shown in the pictures uses printed-circuit boards. There are six different boards. For both 170 and 850 shift the first two are duplicated, making a total of eight.* The boards are laid out by function as follows:

1. Bandpass input and limiter U1.
2. Discriminator and U2.
3. Lowpass filter, slicer, keyer and anti-space.
4. Autostart, standby, meter, delay system.
5. ±12V volt regulated power supply, lamp drivers, meter driver Q10.
6. Loop supply (and room for motor relay).

Room for all parts is included on the boards except for the control switches, meter and the two transformers. The boards are approximately 2¾ inches high by about 6 inches long and plug into 12-pin connectors such as the Amphenol 143-012-01. This is not only neat appearing, it allows the boards to be removed easily for other shifts or to try various

*Printed-circuit boards and complete parts kits for the ST-6 are available from several sources, including Hal Devices, Post Office Box 365, Urbana, Illinois 61801. This firm offers various options and has a brochure available on request. The printed-circuit boards are designed for the 14-pin dual-inline op amps. Hal Devices also have a unique power transformer available that has windings for both the loop voltages *and* plus and minus dc supplies.
For Canadian builders, arrangements have been made with Space Circuits Ltd. to provide printed-circuit boards for Canadian amateurs. Write to Hugh Watt, VE3HY, President, Space Circuits Ltd., 156 Roger Street, Waterloo, Ontario, Canada.

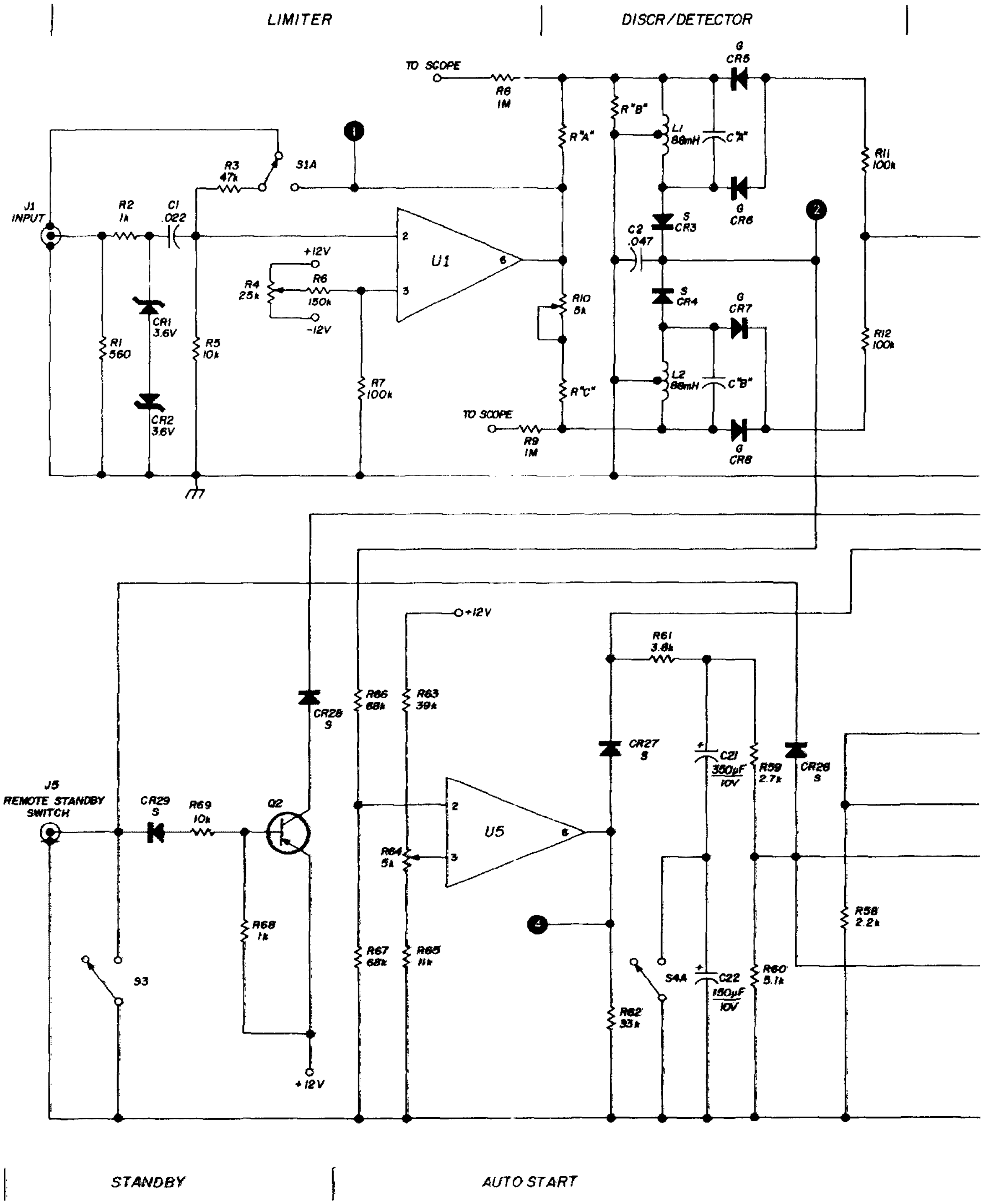
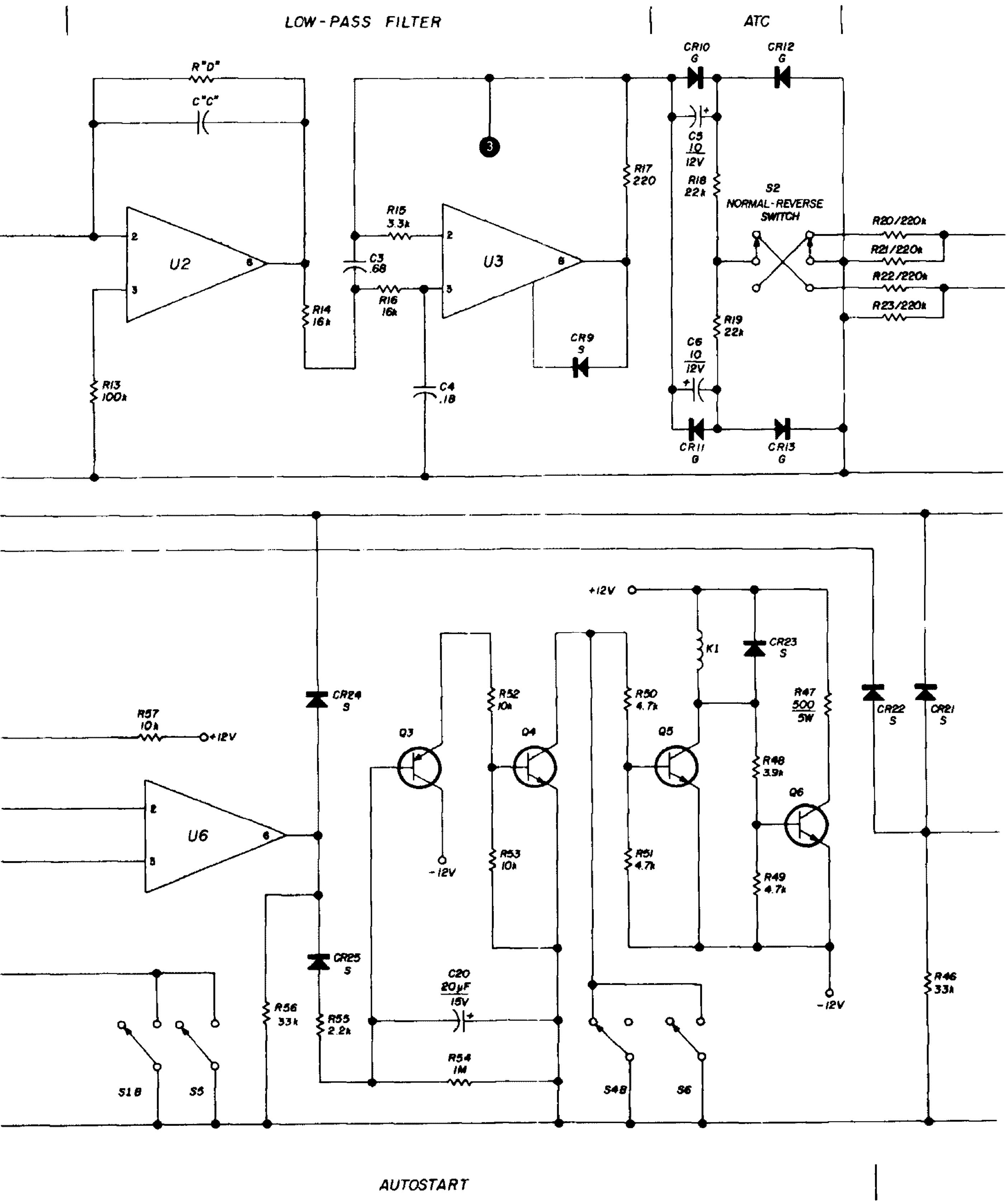


fig. 5. Schematic diagram for the Mainline ST-6 RTTY terminal unit. Contacts on relay K1B open the loop when the motor is off.


LOW-PASS FILTER
ATC
AUTOSTART
R"D"
C"C"
U2
R13 100k
R14 16k
R15 3.3k
R16 16k
C3 .68
C4 .18
U3
R17 220
CR9 S
CR10 G
CR12 G
CR11 G
CR13 G
C5 10 12V
C6 10 12V
R18 22k
R19 22k
S2 NORMAL-REVERSE SWITCH
R20/220k
R21/220k
R22/220k
R23/220k
R57 10k
+12V
U6
CR24 S
CR25 S
R56 33k
R55 2.2k
Q3
-12V
R52 10k
R53 10k
Q4
C20 20μF 15V
R54 1M
R50 4.7k
R51 4.7k
Q5
K1
CR23 S
R47 500 5W
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Q6
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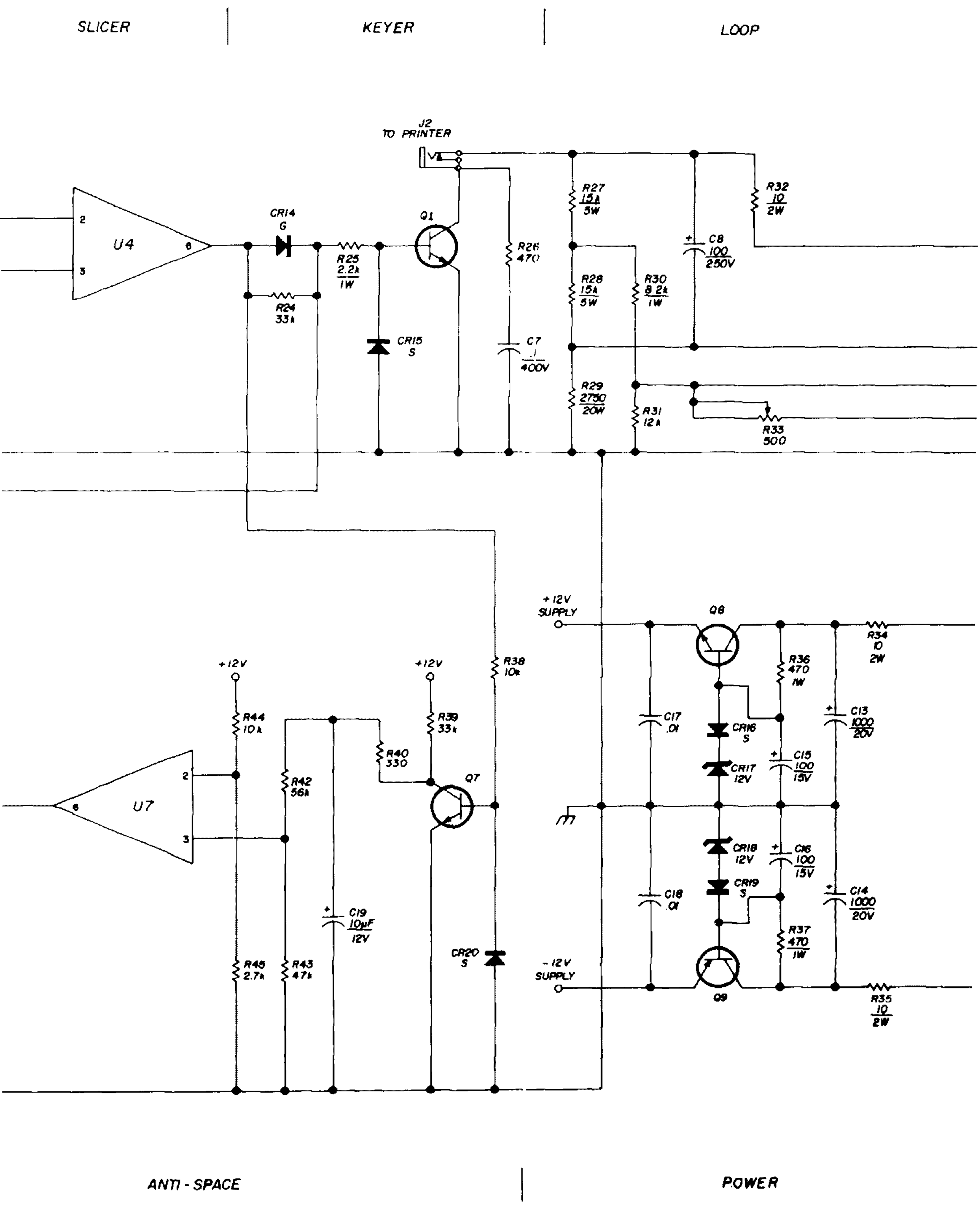
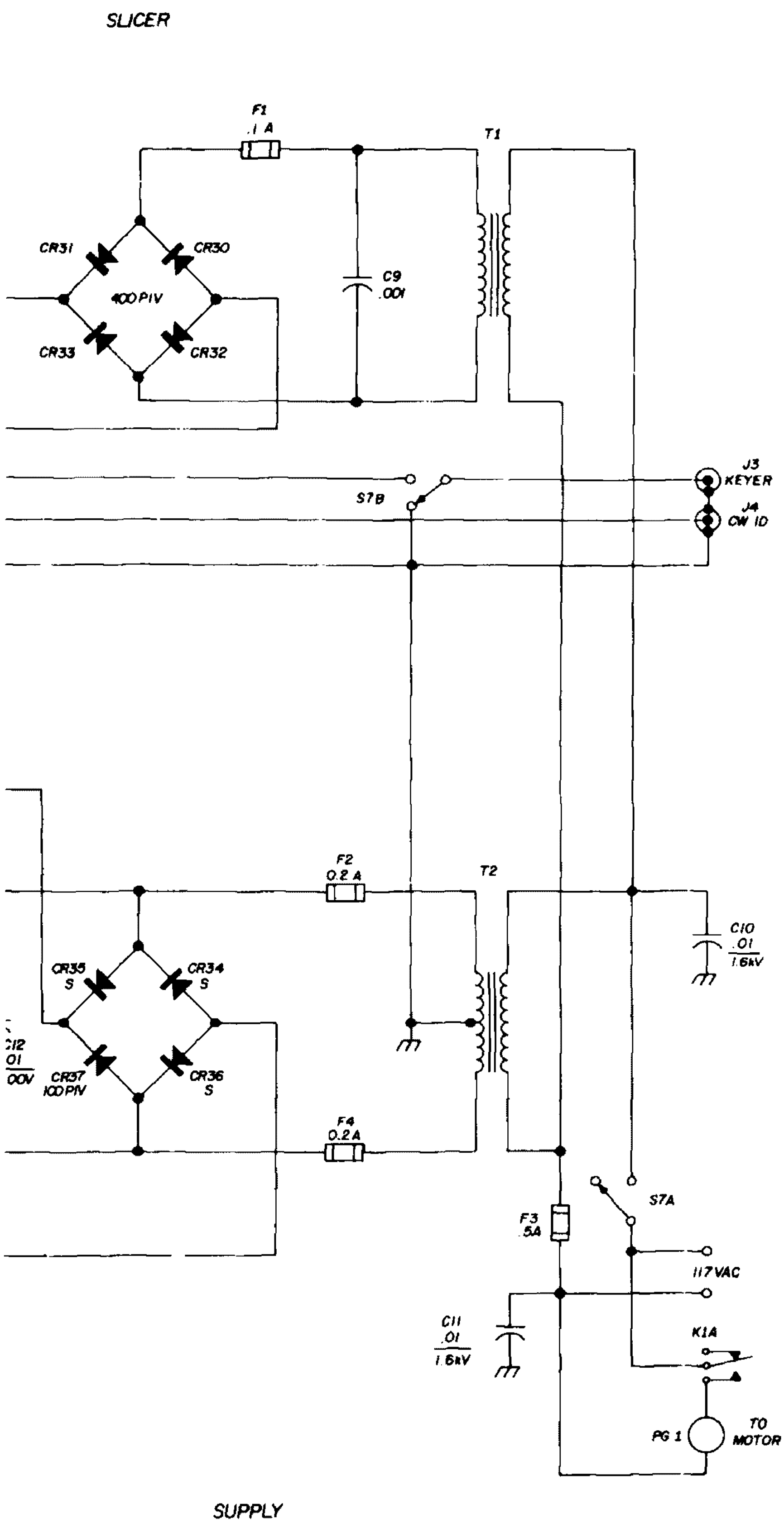
R46 33k
S1B
S5
S4B
S6


SLICER
KEYER
LOOP
J2
TO PRINTER
U4
CR14
G
R24
33k
R25
2.2k
1W
Q1
CR15
S
R26
470
C7
.1
400V
R27
15k
5W
R28
15k
5W
R29
2750
20W
R30
8.2k
1W
R31
12k
C8
100
250V
R32
10
2W
R33
500
+12V
SUPPLY
-12V
SUPPLY
Q8
Q9
R34
10
2W
R35
10
2W
R36
470
1W
R37
470
1W
C13
1000
20V
C14
1000
20V
C15
100
15V
C16
100
15V
C17
.01
C18
.01
CR16
S
CR17
12V
CR18
12V
CR19
S
U7
+12V
R44
10k
R45
2.7k
R42
56k
R43
47k
C19
10µF
12V
R40
330
R39
33k
Q7
R38
10k
CR20
S
ANTI - SPACE
POWER



S1 limiter on-off, also disables autostart

S2 normal-reverse switch, keep wiring short

S3 standby, also turns motor on

S4 fast-slow autostart, fast keeps motor on

S5 autostart off, keeps motor on

S6 manual motor on

S7 power on-off, also ground fsk line when off

T1 isolation transformer, 115 Vac, 0.3 A (Triad N-51X)

T2 filament transformer, 28.6 Vac, 1 A (Triad F-40X)

Diodes marked "G" are 1N270 germanium.

Diodes marked "S" are silicon, 50 piv unless otherwise noted

discriminator values

	170 2125/2295	850 2125/2975
R'A'	6800	4700
R'B'	100k	33k
R'C'	6800	6800
R'D'	270k	180k
C'A'	.068	.068
C'B'	.056	.033
C'C'	.02	.03

Large schematics are available directly from the author for easy-to-follow construction. $1 in USA or Canada, $2 US for airmail elsewhere.

circuit changes the operator may wish to *experiment with.*

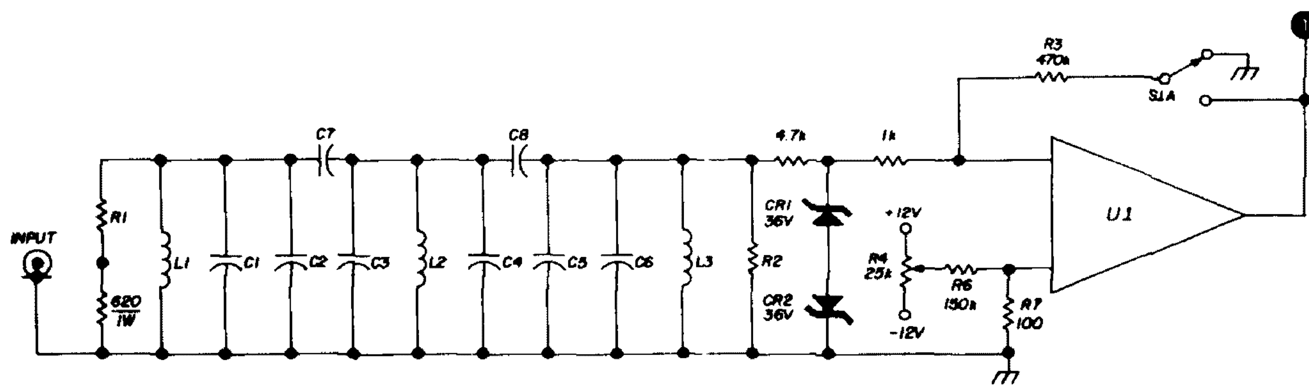
The unit shown in the photos was built into a 10 x 12 x 3 inch chassis, and covered with shelf paper from the local dime store. This makes an attractive way to cover the aluminum chassis and provides an excellent base for rub-on labels.

No shielded lines were used except to the external scope jack. The normal-reverse switch should be as close as possible to board 3 to keep the leads short. The *connectors and switches were all wired* before placing them in the chassis.

are S1, S4, the receive lamp, the standby *lamp, the tuning meter and the* power-on indicator. The latter is superfluous if you have either the meter or the indicator lamps. On the bottom row are S2, S6, S5 and S3, and in the bottom right-hand corner, S7.

On the rear from left-to-right are the remote standby jack, remote scope jack (2-way type), audio input (600 ohms), printer motor, FSK output, fuse, and 120 Vac input.

No cw identification jack was provided as I normally hook that into the fsk line



shift	C1	C2	C3	C4	C5	C6	C7	C8	L1,2,3	R1	R2	tuning frequencies sec1	sec2	sec3
170	.15	.056	.18	—	.15	.056	.022	.022	22 mH	1600	2200	2195	2195	2195
859	.015	.018	.015	.01	.015	.018	.015	.015	88 mH	2700	3300	2400	2300	2400

fig. 6. Optional bandpass input filter.

Adequate spacing is 1" between boards, with board 6 (loop supply) facing the open air with its 2750 ohm, 20 W resistor. This board would be somewhat more than 1" wide when all the parts are mounted.

The printed-circuit boards sold by Halle have transistor connections for Motorola transistors. The Q1, Q8 and Q9 transistors have connections that differ from the others. If you use RCA or GE transistors, be careful to install them properly. About 0.4 watt will be dissipated through Q8 and Q9.

panel layouts

In case you cannot read the printing on the photos, the 170-850 switch is at the far left. Next to it on the upper level

externally, but another jack could be provided as suggested in the schematic.

tuneup

The ST-6 is very simple to align, and once done, should not require any further attention. No pots are located on the front or rear panels. Complete alignment is as follows:

1. Unhook the audio input. Place a dc voltmeter at test point 1 on pin 6 of U1. Adjust the 25k pot on pin 3 of U1 so the meter reads zero. This is a very touchy adjustment due to the high gain of the 709C op amp, so get it as close to zero as is convenient and leave it – this adjusts the offset input balance which allows the op amp to run at its maximum gain level.

2. Either move the dc meter to test point 2 on the discriminator or observe the tuning meter on the front panel. Connect audio to the ST-6 and tune between mark and space signals, adjusting the 5k pot on pin 6 of U1 for equal readings on the meter. This balances the discriminator.

3. If using the tuning meter, with maximum mark indication, adjust the meter sensitivity pot on the emitter of Q10 so the meter reads 70% full scale. Hopefully it will have markings each 0.1 of full scale. Now detune the input frequency to the point where the meter reads 60% of full scale. Place the voltmeter on test point 4 on pin 6 of U5. Adjust the 5k pot on pin 3 of U5 so the voltmeter cannot decide to stay positive or go negative. This sets the autostart sensitivity to approximately ±3 dB bandwidth – about ±45 Hz for 170 shift and about ±100 Hz for 850 shift. If both discriminator boards have been included, set this adjustment in the 170 shift position and take what you get for 850 (850 is much less critical than 170 shift).

If a tuning meter was not incorporated on the front panel, use a voltmeter at test point 2. Tune for maximum mark and note the meter reading. Multiply this by 85%, and retune the input frequency for this new meter reading. Then proceed to adjust the pot on U5 so the voltage at test point 4 cannot decide if it should stay positive or go negative.

tuning the filters

Access to a digital counter makes this job very simple. The Mainline TT/O semi-counter[9] also makes the job very easy. That unit is low cost and hooks to an existing oscilloscope but is quite accurate when compared to anything but a true digital counter. The values of the capacitors in parallel with the toroids in the bandpass input filters and discriminators are only approximate. With toroids that can vary 2 to 3% and capacitors that are ±10%, you can miss the desired frequency by as much as 150 Hz.

When tuning the bandpass input filters, a review of my filter article might be of interest.[10] Basically the procedure is as follows:

1. Disconnect the input and output lines.
2. Disconnect the resistors at input and output of the filter.
3. Short out the center toroid with a piece of wire.
4. Tune the first section.
5. Tune the third section.
6. Remove the shorting wire from the center section and short the first and third toroids to ground.
7. Tune the center section.
8. Remove all shorts.
9. Connect the resistors.
10. Connect the input and output lines.

Room has been provided on the circuit boards for as many as three tuning capacitors for each toroid. This should allow easy, precise tuning without adding or removing turns of wire from the toroid.

The coupling capacitors on the bandpass input filters (C7 and C8) should not be changed from those indicated.

A good method for final checking of discriminator tuning is to put a meter at test point 2 and tune for maximum. Then tune off on the high side to a slightly lower meter reading, note that frequency; tune for the same reading on the other side of center and note that frequency. Average the two to get the best center frequency and change as needed to get the frequency you want.

low tones

This is a subject that is difficult to discuss as part of a demodulator article. Mark tone is customarily 2125; space tone is customarily 2295 for 170 shift or 2975 for 850 shift. Years ago, when these tones were selected good mathematical reasons were used. They are an excellent compromise between channel separation for harmonic suppression and adaptability to communications receivers.

table 2. Components for low-tone discriminators.

		170 1275/1445	850 1275/2125
R'A'	(R1)	2.7k	1.5k
R'B'	(R2)	27k	8.2k
R'C'	(R3)	2.7k	2.2k
R'D'	(R4)	240k	160k
C'A'	(C1)	.18 μF	.18 μF
C'B'	(C2)	.12 + .018 μF	.068 μF
C'C'	(C3)	.022 μF	.033 μF

Many ssb receivers use 2100 Hz audio filters in the i-f and are limited to the range of 300-2400 Hz. Thus, many operators complain that they cannot receive the 2975 tone and assume there must be more appropriate audio tones. Many operators feel that 1275 and 2125 would be suitable. These tones *can* be used and for several years I used them on my 75S-1 receiver and Electrocom demodulator.

However, they are not recommended, particularly if a linear discriminator is used in the demodulator to separate mark and space tones after the limiter.

It is very simple with most receivers to change the bfo frequency with a new crystal so that an audio range of 1400-3500 Hz is obtained instead of 300-2400 Hz. The Collins S-line, the Heathkit SB-line and many others require only a new bfo crystal about 1 kHz lower than that normally used for ssb reception.

The Collins 75A4 and most Drake receivers have passband tuning which automatically allows reception of tones in excess of 3 kHz.

Component values for discriminators for 1275-1445 shift and 1275-2125 shift are shown in **table 2** for those who wouldn't build the ST-6 otherwise. No bandpass input filters have been designed for these shifts nor are any contemplated.

relay circuit

Just a word about an unusual feature – Q6 has a 500-ohm resistor in its collector circuit. Q6's only purpose is to afford a similar current drain on the power supply when the relay is not being used, hence voltage regulation is materially aided. This can be omitted if you are not a purist.

The 47k resistor on U1 was picked for limiterless operation, *without* the bandpass input filter. If using the filter, you will want to change this resistor to a 470k (or whatever value gives adequate meter indication) when running limiterless and using normal audio level on the receiver.

controlled limiting

In limiterless operation if the audio input is more than is needed to reach

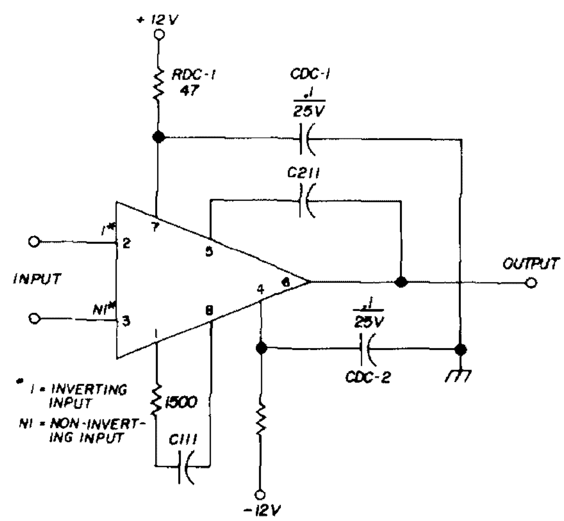


fig. 7. All operational amplifiers used in the ST-6 must be compensated and decoupled as shown here. For U1, C111 is 47 pF and C211 is 3 pF. For integrated circuits U2 through U7, C111 is .005 μF and C211 is 200 pF.

normal meter reading, then the unit goes into modest limiting, sometimes called controlled limiting. This is no longer linear a-m, nor is it hard limiting for fm, but somewhere in between. Some people feel this gives excellent results, but of course you need sharper filters than a linear discriminator offers to take full advantage of any limiterless mode.

cost

The approximate cost using all new parts is $100 for one shift, using vector boards and no PC boards. With both 170 and 850 shift, using the drilled PC boards with connectors, the new-item cost should be around $150.

parts suggestions

Meters may be obtained for as low as

$2.98 from Radio Shack. That is the meter shown on the ST-5. It did a nice job. The one shown on the ST-6 is an imported *Micronta* movement.

The op amps are the 709C types, and Motorola, Fairchild, Texas Instruments, National Semiconductor, and Signetics all make them. The PC boards were made for the round TO-5 type that has eight leads. Prices are dropping constantly. They were $12 each when I first started using them, but at the time this article is being written they are $1 each in small quantities.

The pots used for the PC boards are the 39¢ Mallory MTC-L1 type for vertical

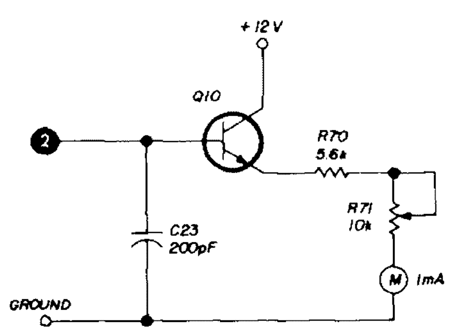


fig. 8. Optional tuning meter for the ST-6.

mounting. IRC type X-201 at 39¢ are similar.

Any 24 Vdc dpdt relay should be adequate. The resistor on the collector of Q6 would be chosen to have the same value as the dc resistance of the relay, such as the Potter and Brumfield KA11DG at $3.90.

The small 0.1 μF 25V capacitors used on each op amp can be the 21¢ Sprague HY-550. Both Centralab and Sprague make 1.6 kV ceramic capacitors used for bypassing the 120 Vac line. The other small value capacitors can be Sprague *Orange Drop* Mylar types, 75 V where obtainable, to keep the size small. Typical price 18¢. The 10-, 20-, 150-, and 350-μF capacitors can be Sprague type 30D. The large 100- and 1000-μF types can be Sprague TVA.

Diodes marked *G* are 1N270 germanium; those marked *S* are silicon with 50 volts PIV except in the loop supply where 400 PIV minimum are needed. The zener diodes can be 400 mW or more, preferably 1 W in the power supply.

Transistors are not critical except in a few places. Q1 should be rated for at least 300 volts; the Motorola MJE-340 is recommended ($1.06). The MJE-340 can also be used in the power supply for Q8. An MJE-370 can be used for Q9 as well as many other types handling 5 W or more – these two being 25 W types. The other transistors can be literally any that are rated for at least 15 volts or more. Those for Q5 and Q6 should be rated at 30 volts or more.

Recommended low-cost transistors for Q2, Q3, Q4, pnp, MPS-3703 (39¢); Q5, Q6, npn, MPS-6565 (52¢); Q1, Q8, npn, MJE-340 ($1.06); Q9, pnp, MJE-370 ($1.07); Q7, Q10, npn, MPS-3394 (27¢); Q11, pnp, MPS-6518 (67¢); and Q12, npn, MPS-3395 (45¢).

The Stancor PA-8421 makes an excellent loop transformer. Other 24 Vct, 0.5 A units should be suitable for the power supply.

The lamps should be low current types, hopefully under 50 mA. Those shown are the Sylvania 925 cartridge indicator lamps using 15 to 20 mA. If you use lamps that pull 80 mA each, change the 10k resistor to the bases of Q11 and Q12 to 4.7k.

The Muralite L-28/40 indicator is ideal for the ST-6. This lamp is rated at 10,000 hours life and may be used with voltages from 12 (25 mA) to 28 (40 mA). They are 39¢ each from: Western Radio, 1415 India Street, San Diego, California 92101; Attn: Gary Pierce Amateur Store Manager.

They are available in violet, red, white, green, blue and amber. Gary suggests you specify "violet leads" as this apparently identifies the L-28/40 bulb, not the color of the lens. These lamps are complete and are normally distributed through the Mura Corp. 355 Great Neck Road, Great Neck, New York 11021. The Sylvania lamps will run around $2.25 each by the time you get the cartridge lamp, lens and holder.

The 709C op amps are now only $1.00

each from Texas Instruments distributors in the 14-pin dual-inline style, and $1.25 for the round 8-pin TO-5 style. By the time this article appears in print it is possible prices could be even lower; as of now the other brands cost about $2.60 each. Texas Instruments uses their own designations for the 709C op amp: SN72709N for the 14-pin dual-inline package and SN72709L for the TO-5 can.

performance

The ST-6 outcopies any other demodulator I have tried. It does not run away from the TT/L-2, but the ST-6 seems to do better *most* of the time. Other operators who have built the ST-6 report similar experiences, and one enthusiast sold both his TT/L-2s and has replaced them with ST-6s. The unit only pulls 5 to 7 watts in standby when the motor is off, so it can run 24 hours a day without affecting the monthly electricity bill. The convenient switching system makes it very simple to use.

The ST-6 is designed for 600-ohm input. Most receivers have such an output available somewhere (on the Collins S-line it is the anti-vox jack). In some receivers, such as the Drake, there is no 600-ohm output. You will throw away about 30-dB potential limiting in the ST-6 if you use the 3.2-ohm speaker line, but small 3.2-ohm to 600-ohm transformers are available for under $1 from most distributors.

other op amps

Some people have asked why I did not use newer, more exotic op amps. Most of the newer devices are internally frequency compensated (such as the 741, for example), and not at all suited for audio use. They have other excellent qualities and could be used in the ST-6 at places other than U1 and U2. Their cost is such that even with external compensation, the 709C is still cheaper. The 709C already has so much inherent potential performance it would be similar to running the TT/L-2 with perhaps 7500 volts on the plate of each tube.

100 speed

For those needing 100 speed, change the two 16k resistors at the input of U2 to 10K, and change the capacitor between pins 2 and 6 of U2 to 60% of its usual value. That is, for normal 60 speed with the 850 shift discriminator for 2125-2975, the normal 0.03-μF capacitor would be changed to 0.018 μF.

The unit would then receive 60, 75 and 100 speed. Do not attempt to switch these items around, just leave it for 100 speed.

conclusion

The ST-6 offers the RTTY enthusiast a high-performance solid-state demodulator. It has features similar to those in the Mainline TT/L-2 but with greater inherent capabilities. It has automatic printer control which is useful for unattended autostart operation as well as for normal QSOs. Printed-circuit boards are available which simplify the construction and make it possible for the home enthusiast to make a commercial-looking unit with only simple tools.

references

1. Irvin M. Hoff, K8DKC, "Getting Started on Radioteletype," *RTTY,* January, 1963, p. 3.
2. Irvin M. Hoff, K8DKC, Keith B. Petersen, W8SDZ, "The Mainline TT/L FSK Demodulator," *QST,* August, 1965, p. 27 (also *RTTY,* November, 1964).
3. Keith B. Petersen, W8SDZ, "The Mainline TT/L-2 FSK Demodulator," *QST,* May, 1969, p. 28 and June, 1969, p. 24 (also *RTTY Journal,* September, 1967).
4. Irvin M. Hoff, W6FFC, "The Mainline ST-3 RTTY Demodulator," *QST,* April, 1970, p. 11.
5. Irvin M. Hoff, W6FFC, "The Mainline ST-5 RTTY Demodulator," *ham radio,* September, 1970, p. 14 (also *RTTY Journal,* May, 1970).
6. Victor D. Poor, "Filters for RTTY," *RTTY,* May, 1964, p.2.
7. W. Mallock, WA8PCK, and T. Lamb, K8ERV, "The Selcal — An RTTY Character Recognizer," *73,* May, 1968, p. 58.
8. Irvin M. Hoff, W6FFC, "AFSK for RTTY," *QST,* February, 1969, p. 11.
9. Irvin M. Hoff, K8DKC, "Checking RTTY Shifts," *QST,* May, 1966, p. 35.
10. Irvin M. Hoff, K8DKC, "High-performance RTTY Filters," *QST,* September, 1966, p. 34.

ham radio

intermittent voice operation of power tubes

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The power **capability** of a transmitting tube is often the subject of long and heated discussions among amateurs (and even among equipment design and tube engineers). In the past, amazing things have been done to power tubes by daring amateurs who seemingly had an inexhaustible supply of replacement tubes at hand.[1] The tube manufacturer looks upon such goings-on with mixed emotions; he's proud his products can take such a beating, but he shudders at the gross overload he knows is taking place, and he has nightmares when he imagines that such tactics are being done by users who may be ignorant of the basic limitations of vacuum tubes. Sometimes that manufacturer may be his own worst enemy. When he speaks of the ruggedness, long life and reliability of his product, he may unintentionally be inviting some eager-beaver to prove the utter conservatism of his remarks.

Up to now, tube ratings have been based upon an absolute system providing "maximum" ratings and "typical" operating conditions for various classes of service, for use below a certain specified frequency. These ratings are designated as *Continuous Commercial Service* (CCS) and *Intermittent Commercial and Amateur Service* (ICAS). The CCS rating may be defined as, "that type of service in which long tube life and reliability of performance under continuous operating conditions are the prime consideration.[2]" The ICAS rating is defined to include the many applications where the "transmitter design factors of minimum size, light weight, and maximum power output are more important than long tube life." The term "intermittent" is used to identify operating conditions in which no operating or "on" period exceeds five minutes and every "on" period is followed by an "off" or standby period of at least the same or greater duration.

These ratings are of cold comfort to today's radio amateur. The first rating applies to high-reliability service (broadcast, military, etc.) wherein off-the-air time is critical or costly; and the second rating, by its very definition, excludes amateur operation in meaningful terms. Neither classification, moreover, applies to ssb or cw operation. Ssb and cw are more properly expressed in terms of peak-to-average power ratio rather than in terms of "on" and "off" periods.

Before new and meaningful ratings are proposed for today's operational modes, it would be prudent to look for a moment at transmitting tube ratings now in use and examine their validity. Contrary to often expressed belief, maximum ratings and typical operating conditions are not arbitrary figures dreamed up by the manufacturer to avoid answering legitimate questions posed by users. On the contrary, they are the result of careful analysis of tube geometry and of prolonged life tests run on typical production tubes with some guaranteed or expected life in mind. Properly understood, the maximum ratings and typical operating conditions can be employed by

the tube user to decided advantage.

transmitting tube ratings

Maximum ratings (or absolute maximum ratings) are those limits within which all tubes of a given type should give satisfactory service and long useful life. Why they are necessary at all and how they are determined are discussed in this article.

The *data sheet* (often suspected of being written by the wise to impress the humble) informs the user of the capabilities and limitations of the tube, both of which are based upon the maximum temperature the elements of the tube can safely withstand for an expected life. Heat, then, is the enemy of unlimited tube life, but heat is the unfortunate consequence of making the tube work. Once the maximum tube capability is determined *a compromise of some kind must be made to establish useful life without exceeding the heat limitations, yet allowing some safety factor for "cockpit troubles."*

maximum plate dissipation

Plate dissipation is limited by the maximum safe temperature of the plate and plate-to-glass (or ceramic) seals of the tube. Generally speaking, the plate will withstand several times its maximum rated dissipation level for a short period of time. Other parts of the tube (glass envelopes, mainly) are greatly affected by the excessive heat radiated by the plate. High level of plate temperature may cause the grid, filament or envelope to become overheated. The grid structure may warp, the filament temperature may rise to an excessively high degree, or the tube envelope may be destroyed. These effects, however, are not instantaneous, and short periods of plate overload do not usually overheat the adjoining tube structure to a damaging extent. However, the user has no way of telling to what degree he can safely exceed the plate dissipation limit, or over what period of time this abuse can take place. The obvious conclusion is that the maximum plate dissipation rating should not be exceeded in continuous operation if long tube life is desired.

maximum plate voltage

The maximum plate voltage point is set at a value above which the internal or external insulators of the tube may arc over, or above which the envelope of a glass tube may be damaged from dielectric losses. Finally, a plate voltage ceiling tends to set a limit to the maximum rf charging current flowing in the plate and screen leads, or plate and grid leads in grounded-grid service. The charging current is a function of the rf plate voltage which, in turn, is a function of the dc plate voltage. Setting a limit on the dc voltage sets a limit on charging current without the difficult task of determining the current directly. This effect depends on frequency and is the reason for the upper frequency limit for maximum ratings.

average dc plate current

The fundamental limit on plate current is the available supply of electrons emitted by the filament or cathode of the tube. The maximum plate-current figure is intended to set a value which may be easily realized throughout the expected life of the tube. If operating conditions are chosen which require that the maximum plate-current limitation to be exceeded at the start of tube life, it may become increasingly difficult to maintain the desired value of plate current as the tube ages. There is a definite relationship between the *maximum instantaneous value* of plate and grid current and the *average dc* (meter reading) *plate current* which differs for each class of tube operation. In linear-amplifier service, for example, most transmitting tubes are run class AB_1, AB_2 (loosely termed class B).* In these cases, the peak plate current is about three times the indicated (average)

*Most class-B linear amplifiers are operated in class AB_2. Class-B operation is defined as cutoff operation with an 180° operating angle of plate current flow. Class AB_2 operation signifies less-than-cutoff condition with more than 180° operating angle.

dc plate current. For long life, the cathode emission should be great enough to provide two or three times the required peak value of plate, plus grid, plus screen current.

The user can quickly determine the allowable average dc plate current in linear service for thoriated tungsten filament-type tubes by merely multiplying the filament watts by a factor of about 5.5. This is a rule-of-thumb number that – over the years – has proven to give a conservative balance between allowable plate current and good tube life. For the 3-500Z, therefore, the filament power is 5 (volts) x 14.5 (amperes) = 72.5 watts. Therefore, allowable maximum average dc plate current for linear amplifier service is 72.5 x 5.5 = 400 milliamperes.

In the case of an indirectly heated cathode the rule-of-thumb is different. Emission from an indirectly heated cathode depends upon the emissive material and the active cathode area, assuming cathode temperature is the proper value. The rule-of-thumb in this case for oxide cathodes is that maximum average dc plate current is approximately 125 milliamperes for each square centimeter of cathode area. For example, the 4X150A tetrode has an active cathode area a little over 2.0 square centimeters and the average dc plate current rating is 250 milliamperes.

long pulse service

In pulse service where the "on" time is small compared to the "off" time, many transmitting tubes can be run to much higher peak power limits than are permissible in continuous service. In continuous service, the maximum voltage and current limitations are set with a safety factor in mind to consider average power dissipated on the tube electrodes. In pulse service, when the tube "rests" for an appreciable time, it is possible to set new guidelines on peak electrode dissipation and maximum ratings, provided the *average* electrode dissipation and maximum temperature ratings are not exceeded.

In pulse service (less than 0.1 second) a thoriated tungsten power tube may have an anode instantaneous peak dissipation capability as high as 100 times the average power capability, and the available filament emission may be as high as 80 milliamperes per watt of filament power. In some cases, the filament voltage has been boosted above normal to obtain emission levels as high as 150 milliamperes per watt with the penalty of greatly reduced tube life.

In the case of the oxide-coated cathode, the peak pulse current is not as clearly defined or as easily generalized as in the case of the thoriated filament tube. A figure of 500 milliamperes peak plate current per watt of heater power is often used for very short pulse service (less than 3 microseconds), and other numbers are available giving pulse plate current in terms of active cathode area.

In *long* pulse service (more than 0.1 second), the rise in temperature of the electrodes rather than the average power during the pulse often becomes the basic tube limitation, and twe maximum capability of the power tube is progressively derated as the pulse length increases. For a radiation cooled tube, a pulse length of 2.5 seconds is often considered equivalent to a continuous duty operation. In the case of an oxide-coated cathode, life tests indicate that a peak-to-average dc plate current ratio of 2.0 for long pulse (0.5 second) is not unrealistic. This corresponds to a duty factor of 0.5.

voice and cw operation

A shadow world exists between continuous duty (CCS) operation, and ICAS operation on the one hand and pulse operation. Amateur voice and cw operations seem to fall into this shadow area. Cw operation may be compared to a form of pulse operation as it defines an "on" and "off" duty cycle wherein the two times are approximately equal. This would represent a duty cycle of fifty percent (0.5) and the pulse (cw) waveform would be nearly square.*

*Note quite true; waveshaping is necessary to some extent to reduce key clicks.

Voice operation, on the other hand, is a different and more complex problem. The voice waveform is not a square pulse; it has a large peak-to-average power ratio with irregular waveform. Normal speech, unclipped, uncompressed or otherwise altered, seems to have a peak-to-average ratio of about 14 dB.[3] Various compression and clipping techniques can reduce this ratio to 3 to 5 dB before severe distortion becomes apparent.[4] Thus, heavily clipped or compressed speech waveforms tend to resemble the cw duty cycle as far as the peak-to-average power ratio is concerned.

It is prudent to expect, therefore, that the power capability of a tube can be safely increased for *Intermittent Voice and CW Service* (IVS service) over the CCS rating provided the maximum element temperature of the electrodes is not exceeded and the cathode (or filament) emission is sufficient to satisfy plate and grid current peaks. In addition, the tube in question should not have an intolerable level of intermodulation distortion when operated in linear service under these enhanced conditions.

This type of intermittent operation is done everyday with the popular sweep tubes used in ssb equipment designed for amateur service. Small soft-glass envelope tubes (i. e., the 6LQ6) are run up to 250 or 300 watts PEP input with no apparent harm *provided* the maximum level of plate dissipation is held within reason, even though the average, long-term dissipation rating in tv service is only 30 watts or so. The user is taking advantage of the intermittent nature of amateur voice operation and the high peak-to-average ratio of the human voice to get more watts per dollar of tube investment. Many amateurs have found, to their regret, an overworked sweep tube tends to overheat and shows extremely short life when a voice clipper/compressor unit is used to bring up the average power of the equipment, or if extended cw operation is used. A moment's reflection upon the heating process in the tube will show the reason for this problem. The tube is being pushed so far that any margin of safety has vanished. Unfortunately, no one has yet been able to miniaturize the watt!

Thus, there's a limit beyond which pushing the transmitting tube becomes uneconomical. It may be well to push an inexpensive sweep tube to 300 watts PEP input, since a tuning error, or other maladjustment won't bankrupt the unlucky user. The owner of a more expensive transmitting tube, however, may well have second thoughts before he blasts his pet power tube. Obviously, some middle ground is called for where the peak-to-average power ratio of ssb and cw operation can afford new and conservative tube ratings more in line with today's usage.

intermittent voice service

In single sideband service, the two plate current values of significance are the single-tone plate current and the two-tone plate current. The ratio of single-tone to two-tone plate current may vary from 1.1/1 to 1.57/1, depending upon the class of operation. Two-tone plate current is useful as the magnitude of intermodulation distortion products may be specified as the reduction in decibels of one product from one tone of a two-equal-tone signal. Precedence exists, therefore, for providing typical operating data for linear amplifier service specifying the dc plate current under two-tone conditions of average plate current and plate current at the peak of the modulation envelope. Such data for the 8122 is shown in table 1.[2]

Based upon such data, extensive life tests have been run at the Eimac Division of Varian to determine if more meaningful operating conditions could be specified for either ssb or cw operating modes. As far as amateur operation is concerned, the limiting mode is cw, where the duty factor is about 0.5. The duty factor for single sideband transmissions with unprocessed speech runs about 0.05 for a 13-dB peak-to-average signal and could rise as high as 0.5 for high levels of speech compression or clipping. A duty factor of 0.5 (peak-to-average ratio of 2.0) for Intermittent Voice Service rating would therefore cover both

table 2. Preliminary operating data for 8873 family of ceramic/metal, zero-bias power triodes.

Cathode: Oxide coated, unipotential	
Warm-up time	60 sec
Heater voltage	6.3 V
Heater current	3.2 A
Direct interelectrode capacitances, grounded-grid connection	
Input	19.5 pF
Output	7.00 pF
Feedback	0.04 pF
Maximum frequency ratings	450 MHz
Operating temperature, maximum, ceramic seals and anode core	250° C
Base	11-pin Special (JEDEC E11-81)
Radio-frequency linear amplifier, cathode driven, class AB_2	
Absolute maximum ratings, to 450 MHz	
Plate voltage	2200 Vdc
Plate current, continuous	250 mA
Plate dissipation	see note 1
Grid dissipation	5 W
Typical operation, Intermittent Voice Service[2] frequencies to 30 MHz	
Peak envelope or modulation crest conditions	
Peak voltage	2000 Vdc
Cathode voltage[3]	8.2 Vdc
Zero-signal plate current	22 mA
Single-tone IVS plate current	500 mA
Two-tone plate current (approximate)	312 mA
Average plate current	250 mA
Single-tone grid current	30 mA
Two-tone grid current	12 mA
Peak rf grid-cathode voltage	67 V
Peak driving power	26 W
Peak power input	1000 W
Single-tone IVS useful output power	587 W
Resonant load impedance	2140 ohms
Intermodulation distortion products:	
3rd Order	-35 dB
5th Order	-36 dB

notes. 1. 8873 is conduction cooled, and plate dissipation depends upon heat-sink cooling. 8874 plate dissipation is 400 watts, and tube has an axial-flow, forced-air cooled anode. 8875 plate dissipation is 300 watts, and tube has a transverse-flow, forced-air cooled anode.
2. Intermittent voice and cw ratings are based upon the maximum voltage and current ratings given for a signal having a peak-to-average power ratio of 2.0 or more. During short periods of adjustment (less than 30 seconds), the average plate current may be as high as the IVS value.
3. Cathode bias is obtained from a zener diode.

the cw and ssb speech-with-processing situations.

The derivation of a different rating from an existing rating may only take place after extensive life tests have been completed to make sure that tube life is not being shortened and that maximum temperature and dissipation limits are not

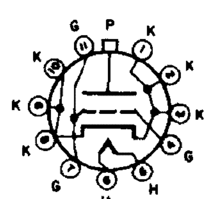

fig. 2. Tube base diagram for the 8873, 8874, 8875 family of zero-bias triodes. Multiple cathode leads keep cathode inductance to a minimum.

being exceeded. Any tube may be limited by grid or screen dissipation level and some may be limited by a plate voltage ceiling, or by available cathode emission. *Each tube type is an individual case, and to jump to conclusions or to interpret data from one type to another is risky and unfounded to say the least.* In all cases the total average current load on the oxide cathode will remain about the same for the new rating as for the average situation.

The *Intermittent Voice and CW (IVS)* rating may be defined as:

> That maximum voltage and current rating given for a signal having a maximum peak to average power ratio of 2.0 or more. During short periods of adjustment (less than 30 seconds), the average plate current may be as high as the IVS value.

In all cases, the IVS rating and "short period of adjustment" are limited by the maximum allowable temperature of the tube anode and seals.

using the ivs rating

The IVS rating is especially attractive to amateur operators as it outlines typical operating parameters for cw and ssb. How are the new ratings used? The following is an example of how an amateur operator can safely and properly tune up for an IVS operating condition with the aid of an inexpensive oscilloscope.

The oscilloscope is necessary for ssb adjustment at first since meter response to a voice waveform may vary from meter to meter and is, in any case, highly irregular and difficult to interpret.

1. The first step to achieve an IVS condition for either ssb or cw is to tune and load the linear amplifier with carrier (single tone) to an IVS rated value of dc plate current (as read on the plate meter), observing maximum "on" time. The amplifier is now ready for IVS CW operation. This could be called the "long-dash" tuning method. An electronic key is handy for this operation.

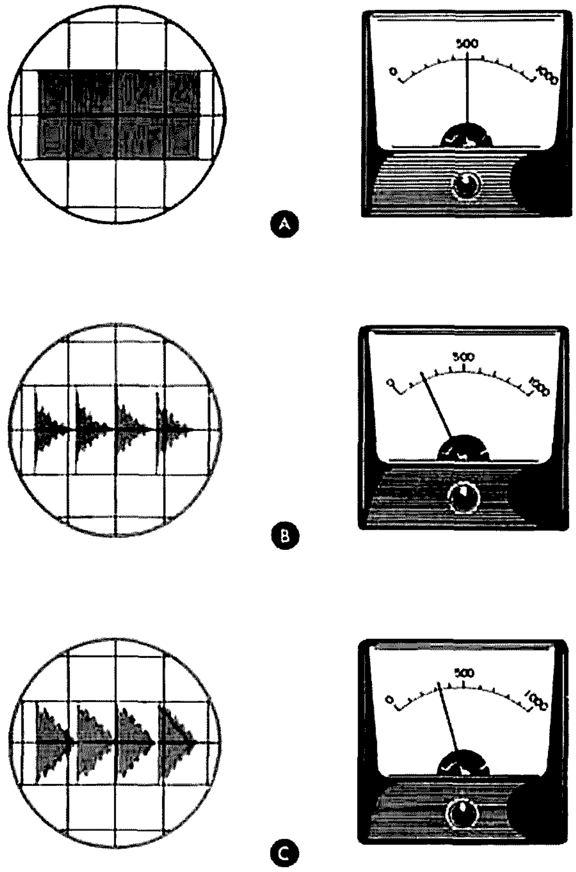

fig. 1. Typical meter readings for IVS operation.

A. With single tone (carrier), the linear amplifier is loaded to maximum IVS plate current (500 mA in this example). Oscilloscope shows carrier pattern. Pattern height is noted as 2 units. Observe short tuneup time.

B. Carrier is removed and voice modulation applied and gradually increased until voice peaks reach carrier height of 2 units as noted in pattern A. Plate meter kicks up to about 200 mA on voicc peaks. No speech processing used.

C. Adding speech processing (clipping or compression). Note that plate meter now kicks up under voice peaks to about 325 mA, but that voice peaks on oscilloscope rise no higher than the single-tone limit of 2 units. However, area under the peaks is greatly enhanced, indicating greater average-to-peak ratio of voice signal. If oscilloscope peaks are greater than 2 units height, with or without voice processing, amplifier is being overdriven, with accompanying splatter and distortion.

2. For ssb observe the rf output pattern on the oscilloscope and note the amplitude for reference.

3. Remove the carrier. Insert audio and slowly increase audio gain so that the instantaneous rf peaks observed on the oscilloscope reach the same maxi-

table 1. Type 8122, linear rf power amplifier service (AB_1). Typical CCS operation at 30 MHz with two-tone modulation.

Plate voltage	2000 Vdc
Grid no. 2 voltage	400 Vdc
Grid no. 1 voltage	-35 Vdc
Zero-signal plate current	100 mA
Plate current	
Peak of envelope	335 mA
Average	250 mA
Grid no. 2 current	
Peak of envelope	10 mA
Average	7 mA
Average grid no. 1 current	0.05 mA
Effective rf load resistance	3050 ohms

mum level as obtained in *step 2* under carrier insertion. The amplifier is now working at the correct IVS level of peak input.

4. Observe the average current peaks on the plate meter for future reference.

In summary, the amplifier is tuned up to IVS condition with single-tone excitation to set the peak signal level. The single tone is removed, and audio is applied so the instantaneous signal peaks reach the same peak level as before, but the peak-to-average level of the intelligence may vary widely, depending upon voice characteristics, degree of speech processing, etc. This is summed up in **fig. 1.**

ivs ratings for the 8873 family of triodes

The new 8873 family of zero-bias triodes is the first to carry the new IVS rating. These ratings are based upon the original design concept of the tube, plus extended life tests where electrode temperatures, cathode emission and power output were carefully monitored. For example, the continuous plate current rating is 250 milliamperes. The cathode area of the 8873 is over 2 square centimeters; this corresponds quite closely to the 125 milliamperes per square centimeter rule-of-thumb stated earlier for an oxide-coated cathode. The life tests showed that a peak dc plate current rating of 500 milliamperes is reasonable at a duty cycle of 0.5 (peak-to-average power ratio of 2.0), corresponding to the IVS philosophy (**table 2**). The various input levels at a given plate voltage may now be established. At 2000 volts, for example, the average plate input is 2000 (volts) x 250 (milliamperes) = 500 watts. This corresponds to key-down service, such as RTTY. The two-tone rating (as in a short two-tone test) is 2000 (volts) x 312 (milliamperes) = 624 watts, average power. The IVS rating for ssb voice or cw is 2000 (volts) x 500 (milliamperes) = 1000 watts peak envelope power. In the case of voice and cw, the average current "load" on the cathode is the same.

Thus, today's power tube may be rated in two different and useful ways. Commonly, it bears the continuous duty (CCS) rating, and occasionally it bears the semi-obsolete ICAS rating. It is hoped that the new IVS rating will find favor in the future as it permits greater operating economy to be achieved in the use of all power-grid tubes.

what about . . .

The immediate question arises, "If this is so, what about the IVS ratings for the 3-500Z or the 8122 or the 4X150A, or whatever?" The present answer to this query is that *each tube type must be examined on its merits* and the outer limits established for any new rating, whether it be pulse, ICAS, or IVS. This is a continual process with most tube manufacturers, and more relevant date of this type will probably be forthcoming over the months.

Thanks to William McAulay, W6KM; Jack Quinn, W6MJG; and Robert Sutherland, W6UOV, for their help in the preparation of this article.

references

1. Perrine, "Thirty-Three Watts per Dollar from a type '52," *QST*, September, 1932.
2. RCA Electron Tube Handbook, HB-3, Volume 1, "Tube Ratings," RCA, Harrison, New Jersey 07029.
3. Magnuski and Firestone, "Comparison of SSB and FM for VHF Mobile Service," *Proceedings of the IRE*, December, 1956.
4. Collins, "Ordinary and Processed Speech in SSB Application," *QST*, January, 1969.

ham radio

modifying the Heath SB-200 amplifier for the new 8873 zero-bias triode

Simple modification of the SB-200 linear to provide increased power dissipation, better frequency stability, and lower drive. Two designs are featured — air cooled and conduction cooled

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The high power capability, moderate cost and compact size of the new 8873 family of zero-bias, ceramic/metal power triodes make them well suited for new design, as well as for retrofit into popular amateur equipment that uses older tubes having restricted power capability and limited frequency range. The well-known Heath SB-200, a one-kilowatt PEP linear amplifier, is a likely candidate. This article covers the modification of this unit to use the new power tubes. The modification provides increased power dissipation,

The Eimac 8875 is a ceramic/metal zero-bias triode with a transverse cooler that provides 300 watts anode dissipation.

better high-frequency stability and lower drive requirements, and (in the case of the 8875) at a lower overall tube replacement cost then the original pair of tubes.

Based upon a study of the SB-200 circuit design, it was decided to try different modifications on two separate amplifiers. The first version uses the 8875, a high-mu power triode having 300 watts anode dissipation and capable of about 1200 watts peak input in Intermittent Voice Service (IVS).[1] The 8875 has five large, round, horizontal anode fins that may be adequately cooled with a small phono-motor fan, the type already in the Heath amplifier. This modification requires a minimum amount of disruption of the existing Heath circuitry.

The second amplifier has a more sophisticated and interesting modification: a conduction-cooled 8873 power triode (electrically equivalent to the 8875) is used with a finned heat sink for proper anode dissipation. The heat sink forms the vertical back wall of the rf enclosure.

This section discusses the first conversion, which provides full input level for the amplifier, with low intermodulation distortion and good tube life. The conduction-cooled version is described in the last part of the article.

the 8875 modification

The 8875 zero-bias power tube is shown in the photo and is capable of 1200 watts PEP input for ssb and 1000 watts when run in IVS service. The tube is about the size of a 4CX250B, has an 11-pin base and uses an inexpensive socket. Cathode and grid connections are brought out to multiple base pins, and, in addition, the grid is terminated in a low-inductance contact ring at the base of the tube which may be used for vhf operation. The anode is intended to be cooled by a horizontal air blast from a small fan. Dissipation is a function of cooling air, and a small phono-motor fan provides about 300 watts dissipation. For RTTY service, the power input level of the amplifier is dropped to about 600 watts. These levels are entirely compatible with the rating of the intermittent-duty power supply of the SB-200 amplifier.

The 8875 is mounted in a horizontal position in the approximate space previously occupied by the two glass tubes as shown in the chassis photo. The 11-pin tube socket is near the center of a small aluminum sub-chassis mounted to the rear wall of the amplifier enclosure. The existing cooling fan, mounted at the

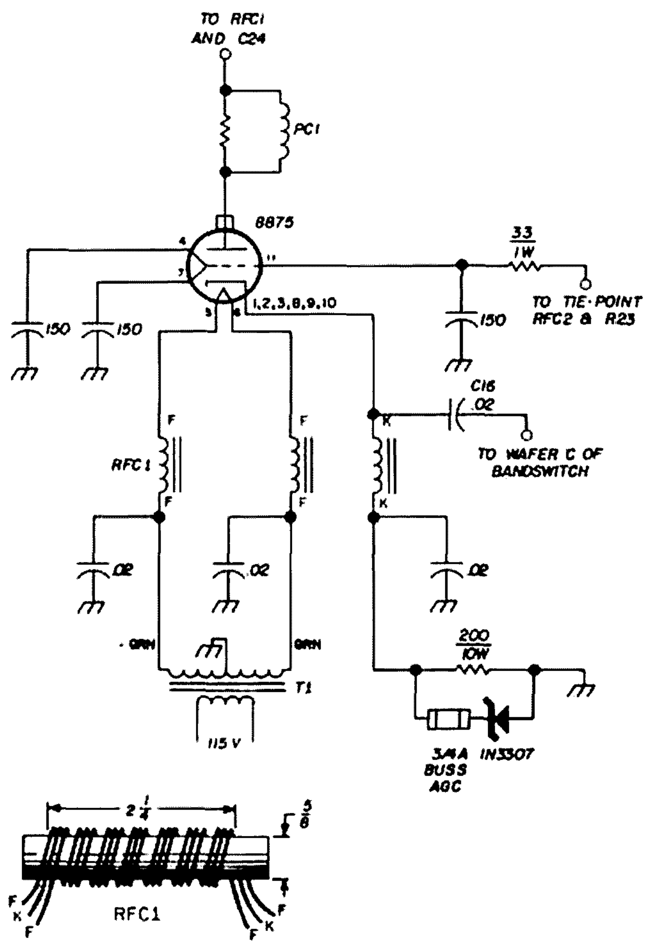


fig. 1. Revised schematic for the SB-200 shows the 8875 zero-bias triode. The 8875 requires 6.3 volts and the filament voltage in the SB-200 is on the high side because of reduced current drain. The original filament-choke windings are removed and three new windings put in their place. The filament windings are two 44" lengths of no. 20 enameled wire; the cathode winding is a 44" length of no. 26 insulated wire. Each winding has a 3" pigtail. Twenty trifilar turns are wound on the ferrite form. The ends are tied with twine and given a drop of epoxy to hold the windings in place. The socket for the 8875 is an E. F. Johnson 124-311-100. The 150-pF grid bypass capacitors are dipped mica units.

bottom of the enclosure, is moved to a new positon in respect to the anode of the 8875.

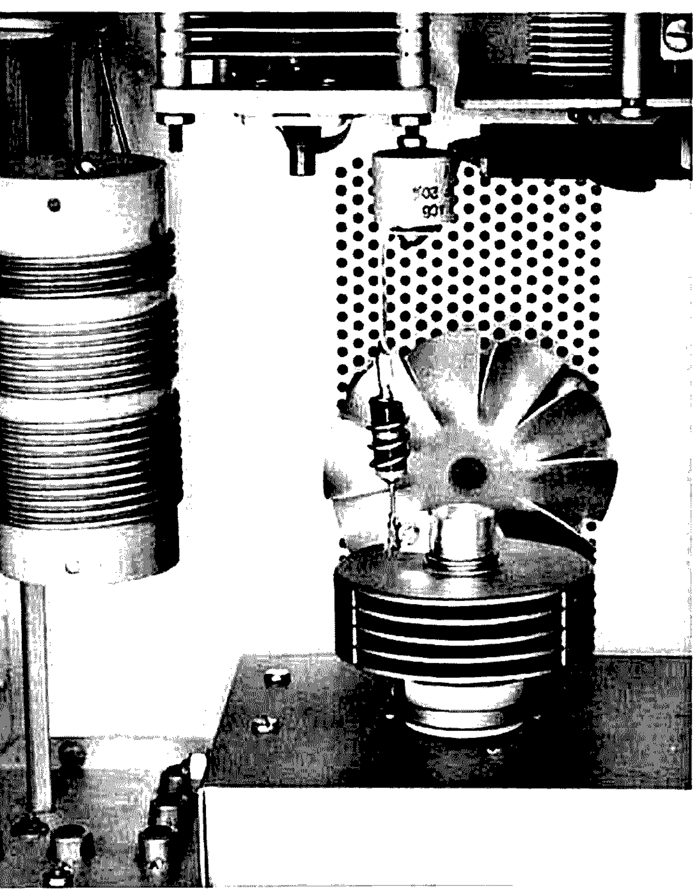
The revised circuit is shown in fig. 1. It uses most of the original components. A new filament choke or dropping resistor is required, as well as zener-diode bias for the 8875. All new components, with the exception of the zener fuse, are mounted within the sub-chassis, as shown in the rear-view photograph.

mechanical modifications

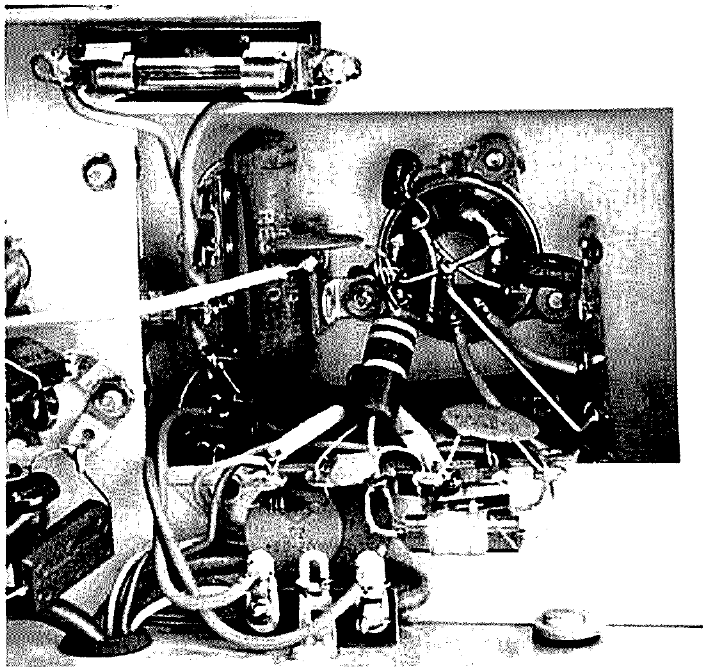
The first step is to remove the components around the existing 4-prong sockets, and then remove the sockets themselves.

Remove the filament choke from tie-point M. Unbolt tie-point AB, leaving the wiring connected. (See pictorial 11, page 39 of the Heath Instruction Manual. Original components are identified by

In the modified Heath SB-200 the 8875 is mounted horizontally in the space formerly occupied by the two glass tubes. Major plate circuit components remain unchanged.



their Heath nomenclature.) The next step is to cut out a rectangular hole on the rear wall of the enclosure as shown in the rear-view photo, with the dimensions



The rear panel of the SB-200 is cut out to allow access to the underside of the new sub-chassis. Untouched cathode coils are to left of the cutout. The zener fuse is at top of cutout with 200-ohm cathode resistor and zener diode mounted in sub-chassis. In this installation an extra cathode choke was used with the original SB-200 filament choke. The cathode choke has 15 μH inductance and 1000 mA current rating (J. W. Miller 4624).

shown in fig. 2. The new sub-chassis for the 8875 is placed over this hole. The sub-chassis is a 4 x 5 x 2-inch Bud AC-1404 chassis cut down to 1-¼" height and held in place with spade lugs and bolts. The tube socket is placed on the sub-chassis as shown in fig. 2.

The new sub-chassis interferes with various bolts holding the rf enclosure to the main chassis deck along the bottom rear edge, and it is necessary to provide clearance for these bolts. Proper clearance is provided by noting position of the bolts and drilling ¼-inch clearance holes in the sub-chassis at the points of interference.

Once the sub-chassis is in position, the

tube is placed in the socket and the phono fan moved until the blades are positioned beneath the anode of the tube as shown in fig. 3.

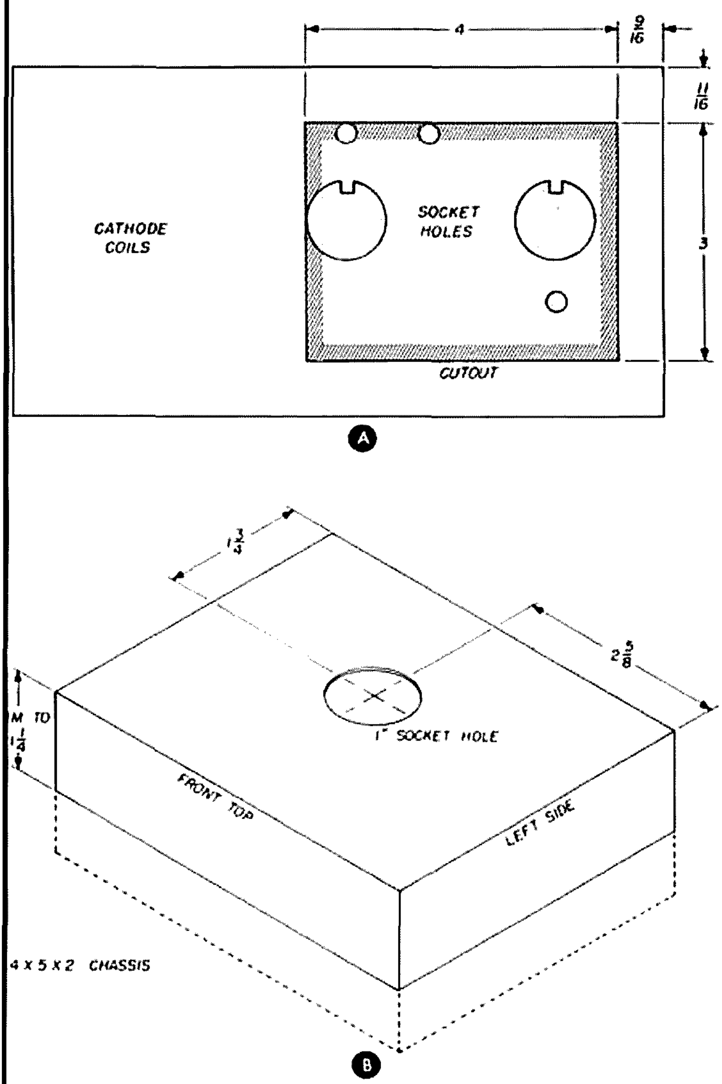


fig. 2. SB-200 chassis modifications to accomodate the 8875. The enclosure cutout for the sub-chassis is shown in A. The modified sub-chassis for the 8875 socket is shown in B.

electrical modifications

The portion of the original schematic of the SB-200 that is revised is shown in fig. 1. The 1N3307 8.2-volt zener diode is bolted firmly to the wall of the sub-chassis, using a thin coating of *Wakefield Thermal Compound* smeared on the zener stud to allow a good thermal bond. The new component layout in the sub-chassis is shown in the rear-view photograph.

Using the new tube, the existing filament voltage of the SB-200 is too high, and it is necessary to drop it slightly to avoid over-volting the tube filament; a 0.3-volt drop is necessary. This may be readily achieved by placing a 0.1-ohm wirewound resistor in series with one filament lead, or the filament choke may be rewound with the proper wire length and size to develop the required voltage drop. Since a cathode rf choke is required, the builder has the option of rewinding the present filament choke and adding a cathode winding as shown in fig. 1 or using the existing choke and adding a cathode rf choke and filament dropping resistor. The latter was done for the first tests, and a new trifilar rf choke was substituted at a later date.

amplifier testing

When the modification is complete, all wiring should be checked and the resistance to ground from the anode clip should be checked. As in the original amplifier, before modification, the resistance should be about 180,000 ohms (the resistance of the filter bleeder resistor, $R_5 - R_{11}$). The amplifier should be connected to the exciter and to a dummy load. Before the amplifier is turned on, the exciter is tuned up, feeding through the unenergized antenna relay of the amplifier. The amplifier is now turned on, and the panel meter should read about +2400 volts in the HV position. Amplifier plate current is zero because of the cut-off bias voltage.

The amplifier controls are set as des-

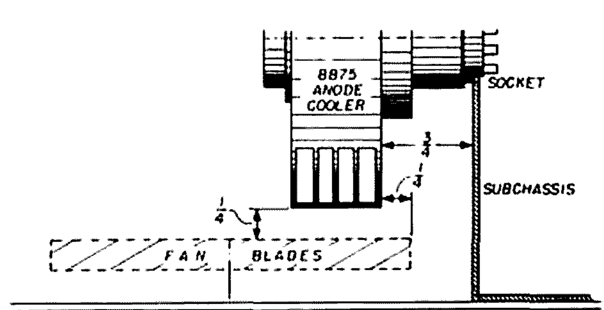
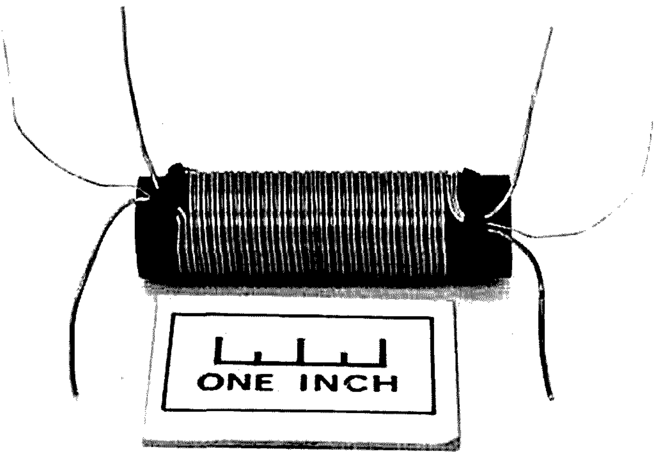


fig. 3. Phono-motor fan installation for the 8875. The motor must be moved so the shaft is in line with the center of the anode, and the tip of the fan blade clears the bottom of the tube by 1/4 inch. The blade should also clear the edge of the anode by 1/4 inch.

cribed in the *Operating Procedure* section of the Heath Manual (page 47), and the amplifier is again turned on. With no driver output, the meter of the amplifier should indicate an idling plate current of about 20 milliamperes. A ninety-second cathode warm up time should be observed before excitation is applied to the



New trifilar filament and cathode choke for the modified SB-200 uses ferrite rod from original unit. Winding details are shown in fig. 1.

8875. The drive level of the exciter is advanced until the plate current rises to about 200 milliamperes. Tuning and load controls are adjusted for maximum power output (minimum plate current) on the amplifier meter. Grid current should be about one division on the meter (25 milliamperes, or less). Caution: *The 8875 is easy to drive; watch out for excessive grid current.*

The amplifier may now be loaded for a maximum plate current of 500 milliamperes, using carrier injection from the exciter. Maximum grid current is 45 milliamperes. This corresponds to a drive level of 55 watts or less. Loading should be done quickly so as to not run excessive IVS plate current for more than 30 seconds or so.

When proper loading with carrier injection is achieved you will find that maximum power output occurs at this point along with the recommended values of plate and grid current. At maximum input the power output (measured with an accurate wattmeter) is between 520 watts (10 meters) and 630 watts (at the lower frequencies). Power gain is about 10 decibels. Under voice conditions, with no speech processing, voice peaks will run about 200 milliamperes on the meter.

Thanks to Merle Parten, K6DC, and Dick Rasor, WA6NXB, for their help and assistance in modifying the amplifier and making measurements on the completed version.

conduction-cooled linear

Modern power tubes such as the 8873-family have the capability of developing anode power dissipation densities (watts per square centimeter) comparable to the power densities in many jet and rocket engines. For this reason, effective cooling techniques are essential for long life and high tube reliability.

Conduction cooling is an efficient system of heat elimination, making use of the *heat source* (power tube or transistor), a heat transmission path (*thermal link*) and a *heat sink,* wherein the heat is removed. Many amateurs have seen transistors with tiny heat sinks on them; far fewer amateurs have observed heat-sink systems capable of dissipating several kilowatts of power. Such large systems exist, and the general design (suitably scaled down) may be adapted for use at amateur power levels. Although common in commercial and military gear, the heat-sink conduction-cooled system is just beginning to appear in amateur equipment (i. e., the Signal-One transceiver).

In the case of a power tube whose anode operates at a high voltage potential, the thermal link must have the dual properties of a thermal conductor and an electric insulator. One of the most practical materials for this task is *Beryllium Oxide* (BeO), an insulative ceramic (refractory) material which has the thermal conductive properties of aluminum.

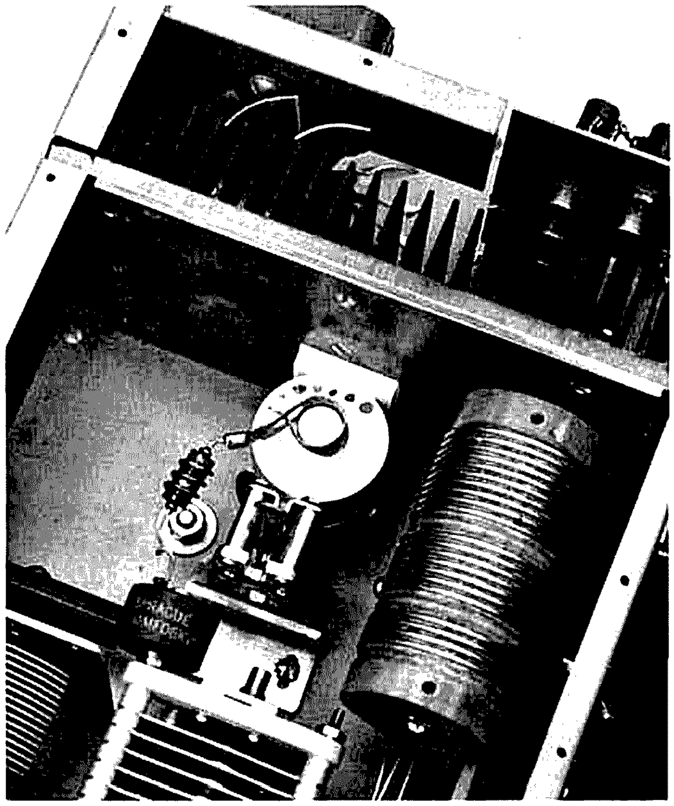
The 8873 zero-bias power triode makes use of a BeO thermal link and external heat sink. The link is detachable,

providing mounting flexibility and reduced tube replacement cost. However, since two thermal interfaces occur with the detachable link (tube anode to link and link to heat sink), attention must be paid to ensure low thermal resistance at these two interfaces if optimum cooling performance is to be achieved.

The heat sink, receiving heat through the thermal link from the tube anode, emits energy in the form of radiant heat. The quantity of heat radiated depends upon the absolute temperature of the sink relative to the surrounding environment and the nature of its surface. A heat sink operating at an elevated temperature compared to its environment will transfer heat to the environment by radiation, convection and conduction, as is done in this case.

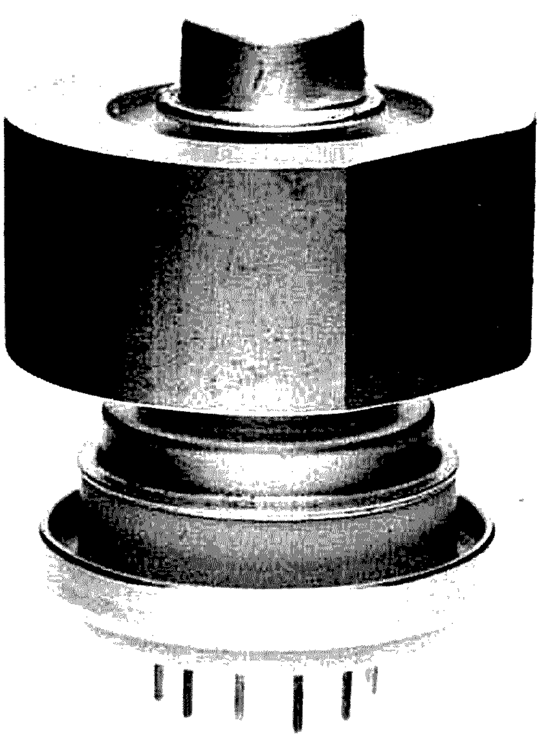
The added output capacitance of the tube supplied by the thermal link and heat sink is typically 6 to 10 pF, and this must be taken into account when vhf tank circuits are concerned.

The heat sink used with power tubes may be liquid or air cooled. In this case



Conduction-cooled 8873 zero-bias triode mounted in the Heath SB-200 linear. The 8873 is held in place with a toggle clamp that presses the anode of the tube against a beryllium-oxide thermal link and finned heat sink. The flat surface of the heat sink is covered with 1/8-inch copper sheet to distribute anode heat evenly. Under normal voice operation heat-sink dissipation provides sufficient cooling. For cw or long-winded voice operation a thermal switch turns on a small phono-motor fan to hold the temperature of the heat sink at a conservative level. When heat-sink temperature drops to normal value, the fan is automatically switched off.

The 8873 conduction-cooled zero-bias triode.



two or three hundred watts of anode dissipation are required so air cooling is feasible.

the 8873

The 8873, like the 8875, is a ceramic/metal, zero-bias triode intended for hf and vhf service up to 450 MHz or so. No air cooling of the base is required if the socket is mounted on a chassis which has

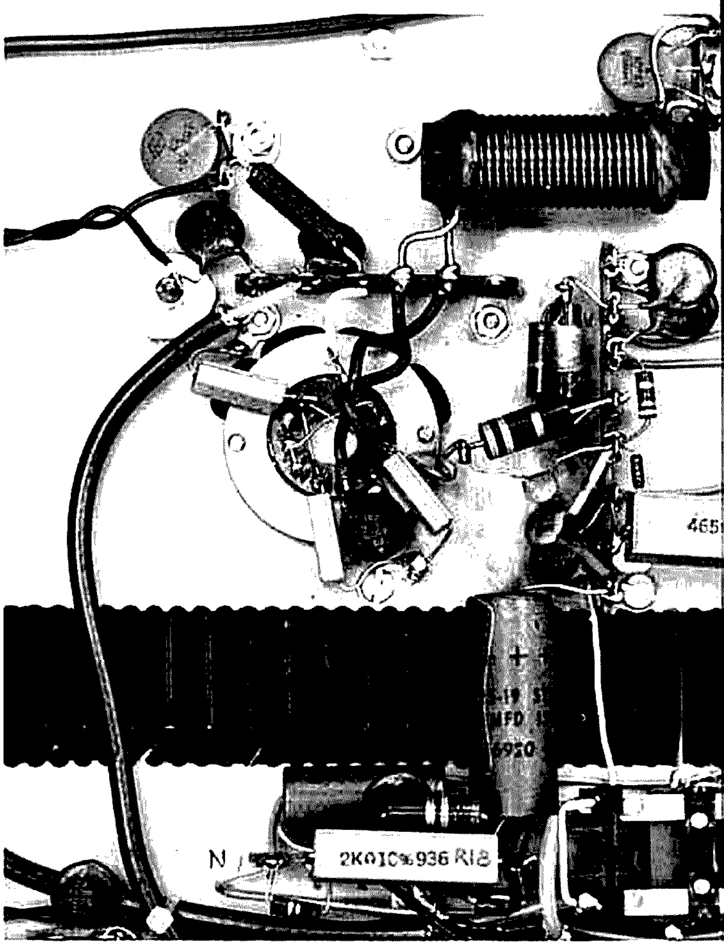
sufficiently low thermal resistance to drain the filament heat away from the stem of the tube.

The 8873 seemed a natural for retrofit in an existing Heath SB-200 amplifier. It was planned that the cabinet and power supply of the SB-200 could be used as a test bed for future experiments (such as a 50 MHz or 144 MHz amplifier) so a new, two-piece aluminum chassis was made. The power supply was rebuilt on one chassis and the amplifier section on another. Both units were then bolted together to resemble the original Heath chassis and shields. An amateur intending to modify his own SB-200 to this design probably would use the Heath metal work as-is.

heatsinking the SB-200 chassis

The 8873 anode is heat sunk to a finned radiator mounted at the rear of the amplifier enclosure. Generally speaking, the modification consists of removing the present tubes, sockets and auxiliary components and reworking the circuit electrically as described in the 8875 modification. Views of the heat sink installation are shown in the photographs. The heat sink measures 7-3/4 x 4-1/4 inches and is mounted to the chassis and side walls about 8-3/4 inches behind the front wall of the enclosure. The 8873 socket mounts in the center of the enclosure, with the center of the socket about 15/16-inch from the smooth surface of the heat sink.

Anode heat flows from the 8873, through a BeO insulating block into the heat sink. Good bonding is essential between these three components in order to hold anode core and seal temperatures below the maximum permitted rating of 250°C. To hold the components firmly together, a DE-STA-CO toggle clamp is mounted in front of the tube. A small ½-inch ceramic insulator is substituted for the rubber nose of the clamp, which presses against the tube and heat sink. While the clamping action takes place, the tube and socket should be free to move. Accordingly, the socket is mounted in a clamp ring so that a slight amount of rotational and lateral movement can be accomplished. The rotational movement is required in order to align the flat surfaces of the tube, thermal link and

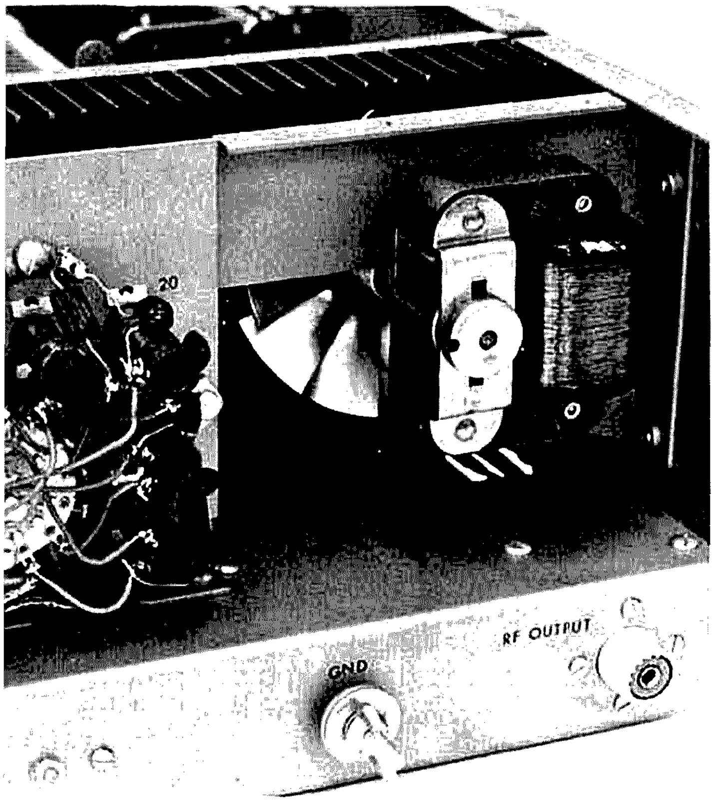


Bottom view of amplifier chassis shows sub-mounted tube socket. Chassis bolt holes are slotted so tube and socket may be moved slightly so the anode is properly seated against the heat sink. The rear of the amplifier chassis has been cut away below the heat sink so cooling air may pass through the fins. Filament choke is at upper right with zener diode mounted on chassis at upper left.

heat sink. The socket is then tightened in position after alignment and clamping takes place.

The heat sink provides about 160 watts of continuous anode dissipation when cooled by normal currents of 20°C air (room temperature). It is possible to raise the dissipation of the sink to about 200 continuous watts by passing cooling

air across it from the small phono-motor fan which was a part of the original SB-200 assembly. The fan was included in this design, along with a thermal switch.



Rear view of conduction-cooled amplifier shows cooling fan and cathode tuned circuits. Chassis beneath the heat sink has been cut away for proper flow of cooling air.

When the temperature of the heat sink approaches a value that indicates high anode temperature, the fan is automatically switched on, increasing the capacity of the sink and protecting it from long-winded rag chewers and marathon talkers.

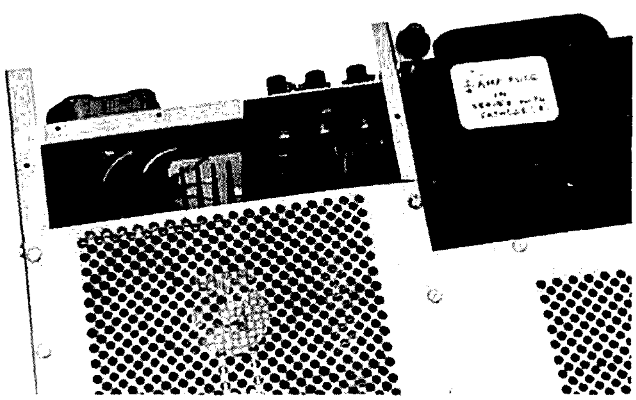
To determine the capacity of the heat sink system, temperature runs were made on the sink and tube anode, with various values of anode dissipation. Heat-sink and anode temperature were measured with temperature sensitive paint, and the thermal switch was moved about on the sink until it switched on when anode temperature reached about 180° C, well below the upper design limit of the tube.

To hold tube base temperature to a safe level, slots were cut in the chassis around the tube socket to allow cooling air from beneath the chassis to flow up and around the tube base (see under-chassis photo).

amplifier operation

Tuning and loading of the amplifier is normal, and follows the procedure outlined in the 8875 description. Under most operating conditions, the heat-sink temperature does not rise to the point at which the cooling fan is actuated, and amplifier operation is completely noiseless, a welcome "sound" these days!

While this unit is considered to be experimental, it points the way to the amplifier design of tomorrow: heat-sunk, noiseless, compact and highly efficient — quite a far cry from the old days of rack mounted gear, heavy, buzzing power supplies and black-crackle panels. How time flies!



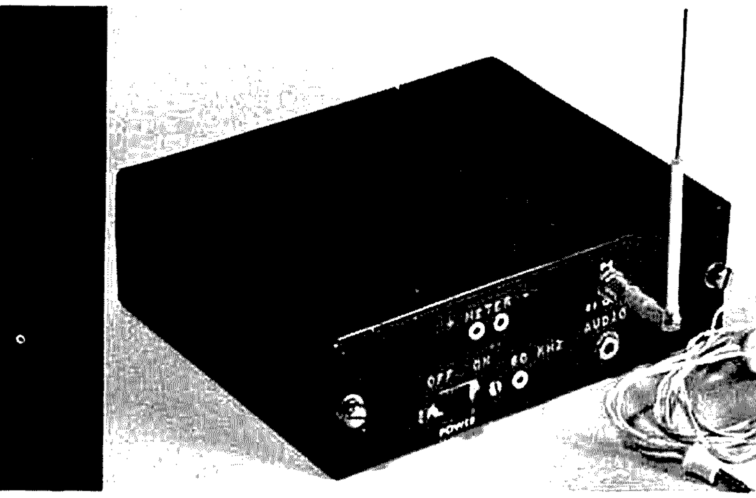
Conduction-cooled amplifier with top shield in place. Shield has been trimmed at rear to allow air to flow over heat sink.

Again, thanks to K6DC and WA6NXB for their assistance in this experimental project.

reference

1. William I. Orr, W6SAI, "Intermittent Voice Operation of Power Tubes," *ham radio*, this issue.

ham radio



photographs by F. C. Lamm, WB4FUJ

two-meter fm frequency meter

A highly accurate heterodyne-type frequency meter that provides crystal-controlled frequency markers on two meters

C. A. Baldwin, W4JAZ 3406 Old Dominion Boulevard, Alexandria, Virginia 22305

On two-meter fm a .005% tolerance crystal can put your transmitter outside the passband of a narrow-band receiver, even if it is working into the same capacitance at which it was calibrated. If the capacitive load is very much different it can even put you closer to the next channel. This is why frequency-shifting capacitors must be adjusted to the crystal in use.

In the local area there are half a dozen hams who have access to laboratory standards and measurement equipment who will help you get your gear on *the* frequency – six *different* ones. These amateurs are performing a very useful service but one wonders where *the* frequency actually is. A fairly accurate answer can be provided by a relatively simple and inexpensive piece of gear.

The unit shown in the photographs measures 2x5x6 inches and weighs 26 ounces complete with batteries. It can be used to adjust a transmitter or receiver to any channel at 60-kHz intervals from 146.94 MHz to below 145 MHz. *Immediately* after calibration it will be within 15 Hz, which is roughly 1/100,000% or one part in 10^7. From day to day, at comfortable shirtsleeve temperatures, the instrument stays within 100 Hz.

how it works

A crystal-controlled oscillator at 49 MHz feeds a buffer-tripler with an output at 147 MHz. This stage is coupled to the emitter of a mixer or modulated amplifier stage. Another crystal-controlled oscillator at 60 kHz with high harmonic output feeds an emitter-follower buffer which drives the base of the mixer, modulating the vhf signal at 60 kHz and its harmonics. The first lower sideband is on 146.94 MHz, and strong signals are available at 60-kHz intervals to below 144 MHz. The output of the mixer feeds a 19-inch whip antenna to radiate signals for receiver calibration.

The antenna and mixer also feed a full-wave diode detector to obtain a beat note between the internally generated signal and a signal at the antenna. A capacitively coupled transistor audio stage provides enough output to a broadcast receiver earpiece to permit adjustment of a transmitter to within about 30 Hz of the internal frequency. Tip jacks on the front of the unit allow connection of a low-range (5-volt) dc voltmeter across the dc detector output. Unless the meter is highly damped, the needle will quiver before the beat note becomes inaudible, and when the meter swings from minimum to maximum and back less than once a second, you have matched the frequencies within 1 Hz.

construction

The frequency meter shown in the photographs was built in a surplus module of uncertain source and purpose which contained most of the parts and circuitry of the calibrator. Where possible the original circuits were used without any attempt to optimize the component values. In the 60 kHz oscillator, it was found necessary to provide a tip jack to read this frequency for calibration checks after the unit was in its case. Since the crystal in this circuit was within 0.1 Hz of the intended frequency, a variable capacitor in series or parallel with the crystal seemed unnecessary. The frequency of the 49 MHz oscillator is adjusted at L1. Substitution of parts or complete circuits of sections of the unit should work equally well.

The circuit is shown in block form in **fig. 1**; a schematic is shown in **fig. 2.** Parts layout is not critical and can follow the dictates of the available housing and the preference of the builder. Some of the features of this unit, however, might deserve consideration.

First, a sturdy chassis is required for high-stability portable gear. All parts must be securely mounted; in this unit the crystals and their sockets are attached to the chassis with epoxy cement. Power to each oscillator is supplied through a dropping resistor with a 6.3-volt zener to ground; a 20% drop in battery voltage can be tolerated without affecting the frequencies. The enclosure was made from flashing copper sheet and permits rf coupling into and out of the unit only at the antenna jack. Audio and meter connections are bypassed at the jacks. Two paralled 9-volt batteries provide long life at the 33 mA load.

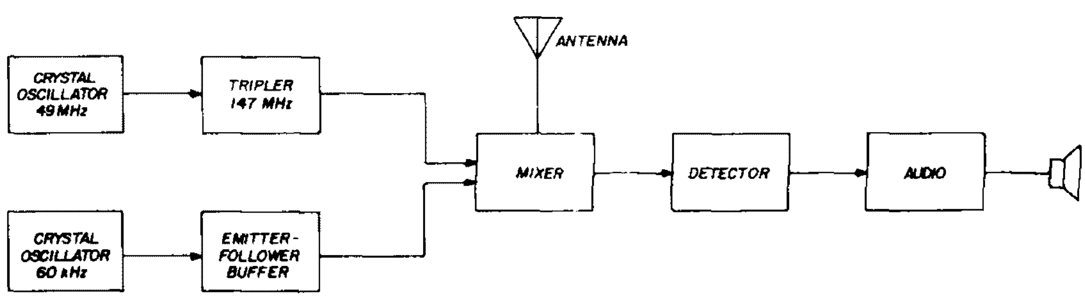


fig. 1. Block diagram of the two-meter fm frequency meter.

calibration

This instrument was calibrated by using an available crystal standard con-

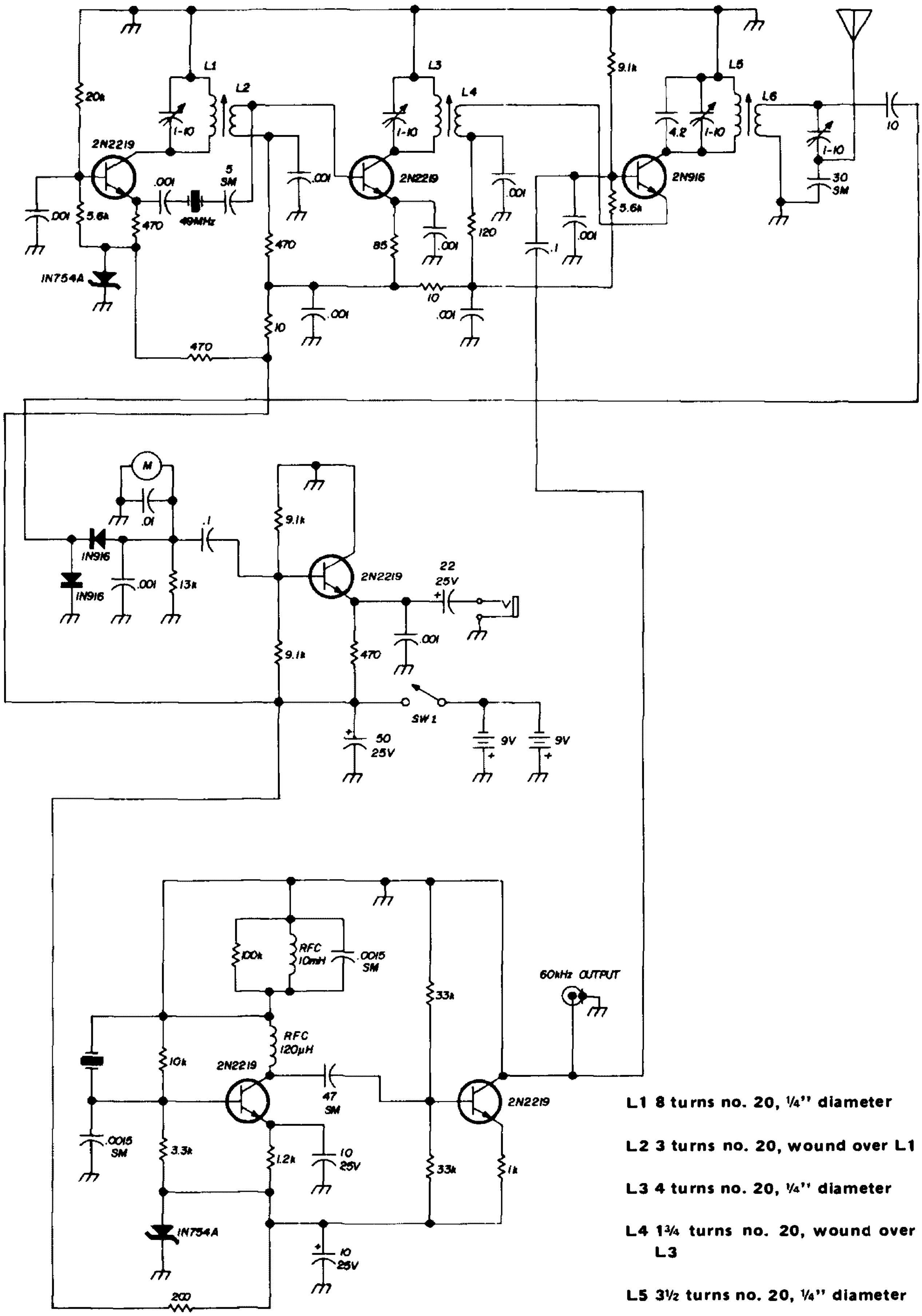


fig. 2. Schematic diagram of the two-meter fm frequency meter. The extensive decoupling circuits and most of the component values were part of the original equipment.

L1 8 turns no. 20, ¼" diameter

L2 3 turns no. 20, wound over L1

L3 4 turns no. 20, ¼" diameter

L4 1¾ turns no. 20, wound over L3

L5 3½ turns no. 20, ¼" diameter

L6 2½ turns no. 20, wound over L5

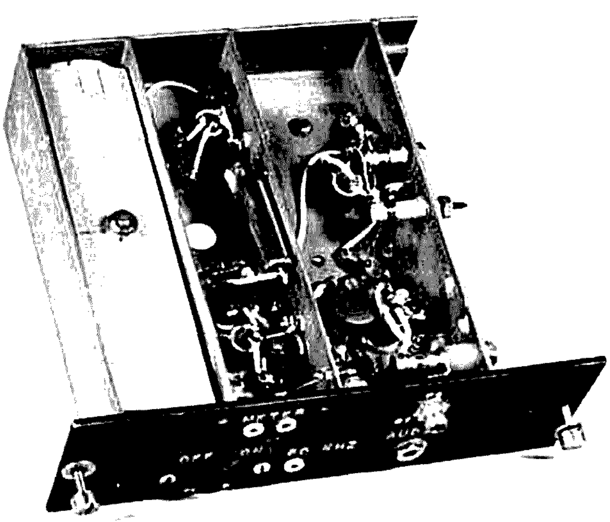
3 turns no. 20, wound over L1

stantly monitored against a WWV broadcast. 10 MHz was used, but where other frequencies are more reliable, the same principles can be applied. This 10 MHz oscillator will hold within 0.5 Hz for the duration of the calibration process. It is followed by a divider to 1 MHz and a multiplier to 50 MHz, both of which feed a mixer. This results in a signal at 49 MHz plus or minus 2.5 Hz. The unit can be adjusted so that its signal, when fed to a receiver along with the 49 MHz standard frequency, will swing the S-meter less than once per second. The cumulative errors at 49 MHz, when tripled, net a maximum error of 10.5 Hz at 147 MHz.

By a similar process, the 60-kHz oscillator can be compared with the WWV signal and adjusted easily to within 0.1 Hz. Since this frequency is multiplied by 1 at 146.94 MHz and by only 20 down to 145.80 MHz, this error does not exceed 2 Hz. It was found necessary to bring this frequency out to a tip jack on the panel for calibration purposes.

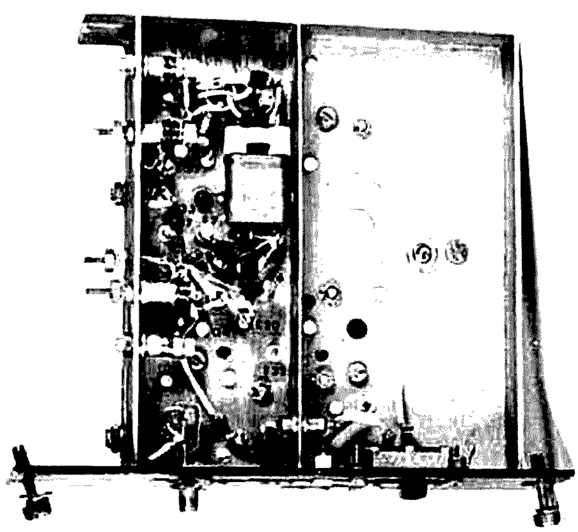
This system results in an error of less than 15 Hz within the intended range of the device. This is not as good as the best laboratory counters, but is cheaper, more rugged, portable, battery powered, and always available.

Top view of chassis. The 60-kHz oscillator is located next to the batteries; buffer is at rear. On the right, front to back, are the antenna coupling capacitor, detector and audio stages, and tripler.



use

The simple frequency meter is used just like the LM, BC 221, or other heterodyne frequency meters with an audio detector, except that a low-range, high-resistance dc voltmeter can be used to monitor the beat note at less than audio frequency. The signal from the unit can be picked up on the receiver under

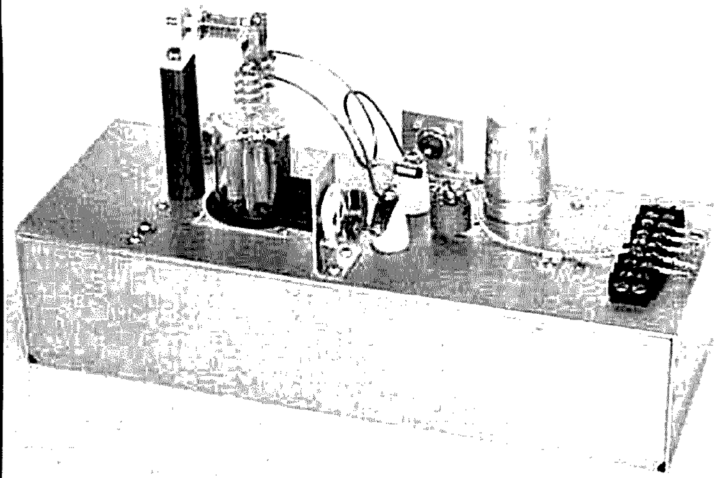


Bottom view. The 49-MHz oscillator is at rear, mixer in front.

test, and the receiver's oscillator adjusted until its discriminator voltage is zero, or as otherwise directed for that receiver.

To adjust a transmitter frequency, the unit must be within about 20 feet of the transmitting antenna for powers of ten watts or above. The transmitter frequency is adjusted for zero beat as heard in the earpiece or shown on the meter. In practice, little equipment has been encountered which can be adjusted below the beat note range without meticulous care. But this is much better than most of the ham transmitters you are likely to hear. Also, it is comforting to know that the error is mostly in the transmitter or receiver adjustment, and not in the test equipment.

ham radio



an rf power amplifier for 220 MHz

Have you been avoiding vhf operation because of construction difficulties? Then meet the "220 Lady"

M. O. Beck, WB6DJV, 1637 Hood, Wichita, Kansas 67203

The apparatus shown in the photographs accompanying this article is a class-C rf power amplifier for the 220-MHz amateur band. It was designed and built to disprove the remarks made by hams who claim that operation on 220 MHz presents too many technical difficulties. The classic remark is, "220 MHz is just too hard to get on." Well, all that is changed now.

The amplifier described here has features that should appeal even to the most skeptical vhf critic. Standard parts are used, and if my layout and construction suggestions are followed, the amplifier will give a good account of itself with a minimum of debugging.

After a considerable amount of thought on the subject of why many hams shy away from building vhf amplifiers, I arrived at the real reason: neutralization. Since this is the main objection, I decided to put it out of the picture once

and for all by designing an amplifier that requires absolutely no neutralization whatsoever. I further decided that the amplifier should require such minimal drive that the driver could be a multiplier (again, no neutralization).

I dubbed this amplifier the "220 Lady," simply because it behaves like a lady – in the classic sense, that is! From the moment it was first fired up, it was a perfect example of what a good amplifier should be: tame, cool running, and efficient. In fact, the night it was completed this amplifier was the subject of a lecture given at the Wichita Amateur Radio Club.

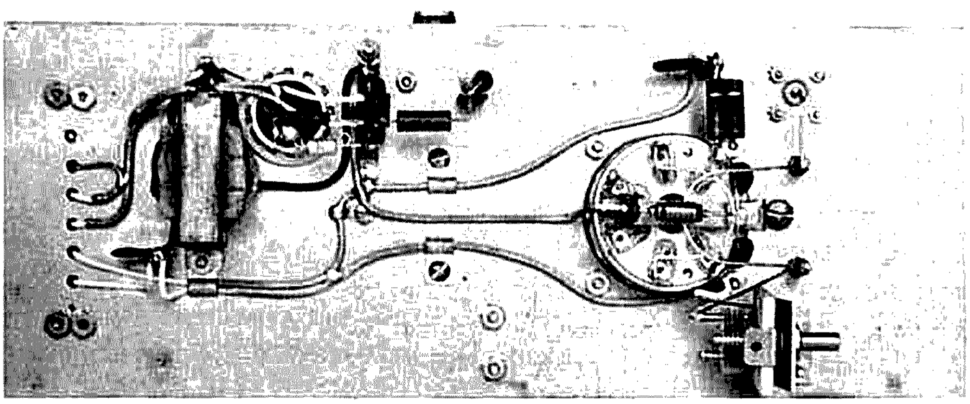
Other than praise, there's little else to be said about the amplifier, as the schematic (fig. 1) and the excellent photographs (taken by a friend, Mike McConnel) just about tell the whole story.

publication for beginner and experienced vhfer alike.[1]

I'd like to emphasize one important point. It's strongly recommended that my design and layout be followed exactly. Parts layout, lead dress, and mechanical details have been very carefully worked out. The shielded wire (see photos) is an absolute necessity. Although some variation could be made in parts layout, I don't recommend it. The grid and plate tank circuits are rather old fashioned, but that's probably why the amplifier works so well. This construction method (fig. 2) is tried and true. If the amplifier "takes off" then you have made some mistake in wiring or have side-stepped my layout.

bias requirements

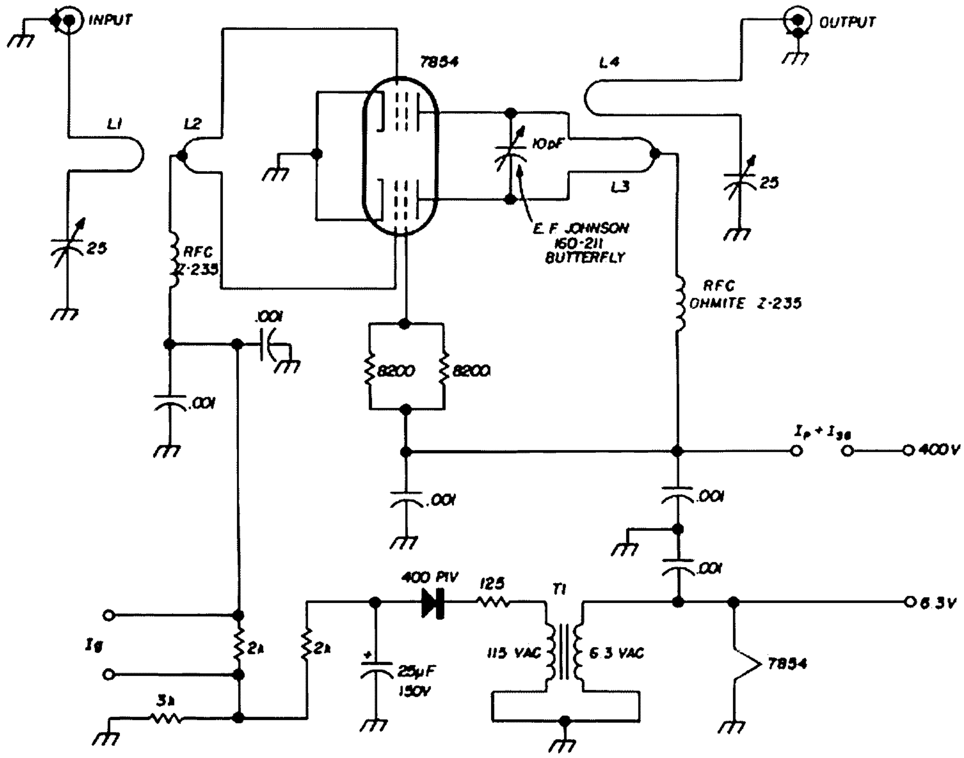
I hope you'll put the little bias supply where I did – under the chassis where it belongs! As to setting the bias voltage, I've included photos and the schematic of the supply I use. It may differ a bit from your idea of what class C should be, but before you knock it – try it. I strongly advise against keying the amplifier in the cathode, as instability could result. We're talking about class-C operation, so key the transmitter where it should be keyed – in the exciter.



Under-chassis view. Shielded wiring is necessary for circuit isolation at this frequency. Note simple bias supply, which doubles as tube-heater source.

construction

The fact that you've read this far indicates you're interested in getting on 220 MHz. I'll assume you've had some experience in construction and wiring techniques at the vhf-uhf level. If not, all you have to do is read one of the many articles published on the subject. The vhf-uhf manual by G6JP is an excellent



Top-chassis view of the amplifier. Parts layout should be close to that shown for trouble-free results.

phone operation

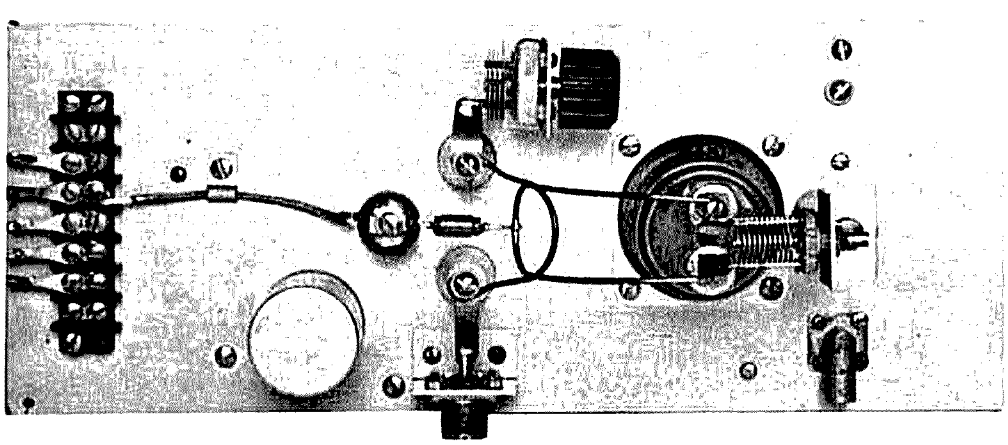
You want to modulate the rig? By all means, go ahead and plate modulate; remember, the amplifier is operating at class C. Interested in ssb? Just decrease the bias and increase the screen voltage, which should be obtained from a separate source, of course, Your scope will tell you when you have the correct bias adjustments.

tube considerations

If you have a 5894 it may be substituted for the 7854 but this may present neutralization problems which we're trying to avoid. Using the 7854, I obtained rated output with a type 6360 tube as a driver-tripler. Drive power is no problem; only a watt or two is required.

The amplifier should not be operated at dc inputs exceeding 50-60 watts. Initial

fig. 1. Schematic of the "220 Lady." Neutralization has been eliminated in this amplifier. Full dc input of 50 watts requires only one or two watts of drive.



circuit adjustments should be made with the aid of a grid-dip oscillator. These dual tetrodes (7854, etc.) don't take kindly to excessive out-of-resonance maltreatment.*

conclusions

This amplifier is just about the ultimate from the standpoint of cost and construction simplicity, but don't be deceived – it's a real performer. The amplifier is designed for cw and really does a good job in this mode.

This article is a cart-before-the-horse thing. I told Jim Fisk I had an article on the exciter for this amplifier, but second thoughts led me to rebuild it for appearance as well as efficiency. So maybe this amplifier will at least sooth the editor as well as convince a few more fellows that it really isn't "hard to get on 220 MHz."

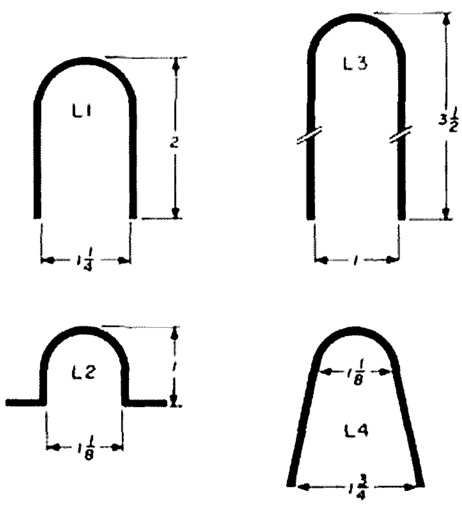


fig. 2. Dimensions for plate and grid inductances. Material can be small-diameter copper tubing or large-diameter copper wire.

reference

1. G. R. Jessop, C. Eng., MIERE, G6JP, "Vhf-uhf Manual," Comtec Book Division, Box 592, Amherst, New Hampshire 03031 ($3.75 ppd).

ham radio

*Although not shown in the schematic, good insurance against tube failure is a screen-grid protective circuit. Several schemes are shown in the ARRL handbook. editor.

88 and 73

The numerals 88 and 73 have been a tradition in communications language for almost 120 years. The older of the two, 73, appeared in 1853, meaning "My love to you." In 1857, the first official definition made it a "fraternal greeting between operators." Two years later, 1859, Western Union made "73" a part of their "92 Code" to indicate "Accept my compliments." The final change came in 1895, when "73" meant "Best Regards" for the telegraph, and later for radio operators.

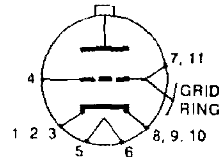
"88" never received the formality of an official listing until it was adopted as one of the "Ham Abbreviations." It had been one of the telegraph operators' traditional terms since well before the turn of the Century. During the First World War, "88" was used by the U. S. Army Signal Corps, again strictly as an operator's abbreviation in unofficial communications. At the close of WWI, "88" achieved official status as a part of amateur radio terminology: "love and kisses."

Louise Ramsey Moreau,
WB6BBO/W3WRE

More Details? CHECK–OFF Page 94

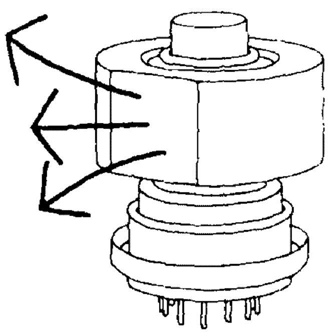
1. Up to 1000 watts PEP input per tube to 432 MHz. Typically, 2000 volts at peak dc plate current of 500 milliamperes.
2. Indirectly heated cathode. No expensive filament choke needed for grounded grid HF operation as filament is electrically isolated from cathode.
3. Easy to drive. Extremely low grid interception plus EIMAC's exclusive self-focusing cathode combine to provide high power gain, high overall efficiency and low, low intermodulation distortion.
4. May be driven to full rated input with solid state driver in many cases.
5. Inexpensive ($1.25) socket. No air ducting required. No expensive glass chimney. 11-pin low inductance base and VHF grid ring provide excellent intra-stage isolation.
6. Heat sink cooling available to full anode dissipation.
7. Tubes may be mounted in any position, greatly aiding circuit layout.
8. And, of course, no bulky screen or bias supply needed!

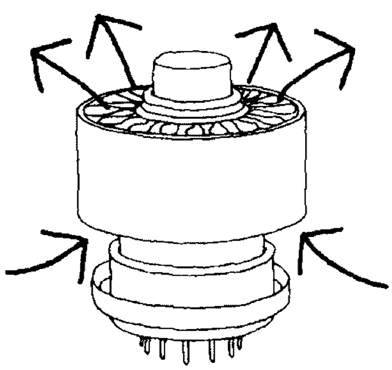
8873 FAMILY BASE CONNECTIONS

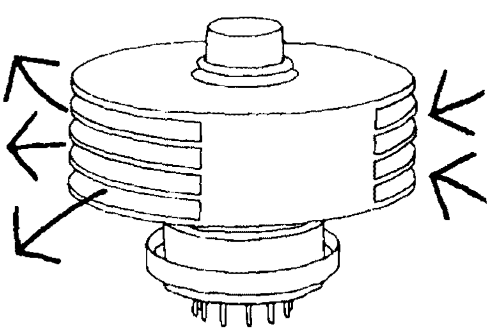


JEDEC #E11-81

Another example of EIMAC's ability to provide tomorrow's tubes today!







Here are the numbers to prove it:

The EIMAC 8873 is a grounded grid triode designed for conduction cooling to a heat sink. See this 1971 design in tomorrow's ham gear.

The EIMAC 8874 is electrically equivalent to the 8873 but has a 400 watt anode designed for axial-flow, forced air cooling.

The EIMAC 8875 is also similar to the 8873 with a 300 watt anode designed for transverse air cooling with a quiet and inexpensive "phono motor" fan.

A pair of any of these exceptional triodes will fit in the palm of your hand. And will provide you a full 2000 watts PEP input for voice. They're rated for continuous RTTY service, too. We've built a single tube kilowatt 432 MHz stripline amplifier. It'll run 50% overall efficiency with 25 watts drive. More than adequate for successful moonbounce operation!

14 MHz DX operation . . . RTTY . . . moonbounce at 432 MHz . . . widely separated goals, but all met by one family of rugged, ceramic/metal grounded grid triodes, by EIMAC.

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Plate Voltage	Zero-Signal Plate Current	Single-Tone IVS Plate Current	Useful Power Output
2000 V	22 mA	500 mA	587 W
Peak Drive Power	**Cathode Voltage**	**Resonant Load Impedance**	**Intermodulation Distortion Products**
26 W	+8.2 V	2140Ω	3rd order: −35dB 5th order: −36dB

Write to EIMAC for full details and suggested circuitry for this new family of tubes. The 8873, 8874 and 8875. And 8873 to you. EIMAC, 301 Industrial Way, San Carlos, California 94070. Phone: (415) 592-1221.

division
varian

inexpensive swr indicator

A simple, easily constructed device that indicates behavior of your feedline at any point

J. A. Houser, WB2GQY, 23 Washington Street, Rensselaer, New York 12144

The fact that this standing-wave indicator is inexpensive in no way detracts from its usefulness as a measuring instrument in the ham shack. It's not an in-line device and is not connected to the transmitter. Therefore it contributes no insertion loss and doesn't disturb the circuit with which it's used. It is simply an rf-current loop connected to a half-wave rectifier that actuates a sensitive dc microammeter.

antenna feedlines

I believe it's self-evident that there's no such thing as a perfect antenna. This leaves the ham with a choice of alternative compromises. He can erect a dipole antenna, which seems to be the most popular on the four lower-frequency amateur bands. Then comes the decision as to the best method of feeding the dipole. More and more hams are using RG-8/U 50-ohm coaxial cable. The reasons are low cost, trouble-free service, and high efficiency.

The low impedance of RG-8/U dictates a current-fed system. It's well known that the feedpoint impedance of a half-wave dipole is closer to 70 than 50 ohms. This means that a mismatch will exist between the feedline and the antenna, which is a starting point for a whole series of loss parameters.

cut-and-try matching

One of the usual ways of compensating for the mismatch between 50-ohm line and the antenna feed point impedance is to start cutting the line, which may partly reduce mismatch loss to a minimum. Another method is to use some type of matching device between transmitter and antenna so that the transmitter is looking into the terminating impedance for which it was designed. This involves added expense, so the average ham uses the cut-and-try approach until the transmission line electrical length is close to a half-wavelength or multiple thereof at the operating frequency. When this condition is met, the transmitter will load properly.

All the cut-and-try matching method does is improve the transmitter-to-line match; it does not improve the impedance match between transmission line and antenna. Transmission lines with a characteristic impedance of 70 to 75 ohms present a better match to a dipole, but this line is more expensive for the maximum legal power. Therefore, most hams use the more inexpensive 50-ohm cable and live with the consequent impedance mismatch.

Any mismatch — at the antenna or transmitter or both — can be indicated to a varied degree by an swr meter, which is generally inserted between transmitter and the load. When the swr meter indicates a 1:1 ratio, this means that the transmitter is presented with a load whose impedance is equal to that of the

transmitter output circuit. It *does not* necessarily indicate that the transmission line is matched to the antenna.

matching at both ends

How then, you might well ask, can one tell how closely the feed line is matched at the transmitter *and* the antenna? Merely by building and using the swr indicator described here. Because this instrument will indicate the standing wave ratio at any point, and therefore at *all* points along the feedline, it is much more useful than the ordinary swr bridge/indicator usually employed.

If you doubt what I'm saying, just take that swr bridge you've purchased and insert it into your transmission line at various distances from your transmitter by adding lengths of line, and you'll see immediately what I'm trying to put across. You'll find that the swr meter reads differently for each length of cable added. If it doesn't, pat yourself on the back, because you have a near-perfect impedance match between your rig and the antenna.

construction

The swr indicator is simple to build, as the photo shows. Total cash outlay for parts will be somewhere between $5.00 and about $14.00, depending on how much you want to spend for the most expensive component – the microammeter. A complete parts list is provided in table 1.

The coil is constructed with no. 14 dcc wire. Six turns are wound on a ¾-inch-diameter dowel, with one layer of Saran wrap between wire and form to make coil removal easy. A coat of polystyrene dope is then applied to the coil to make it rigid.

A 20-microamp meter makes a sufficiently sensitive indicator, which will indicate currents as low as 1 microamp – sufficient for very low-powered transmitters.

I used a 1N34A diode rectifier, as it exhibits little frequency sensitivity from the lowest ham band to 250 MHz. Sufficient indication was obtained on all bands, so no tuned circuit was used. However, the sensitivity can be increased for one particular frequency by connecting a variable capacitor across the coil.

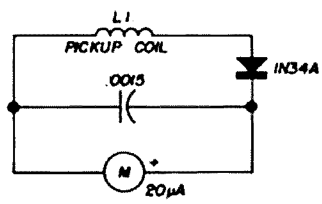
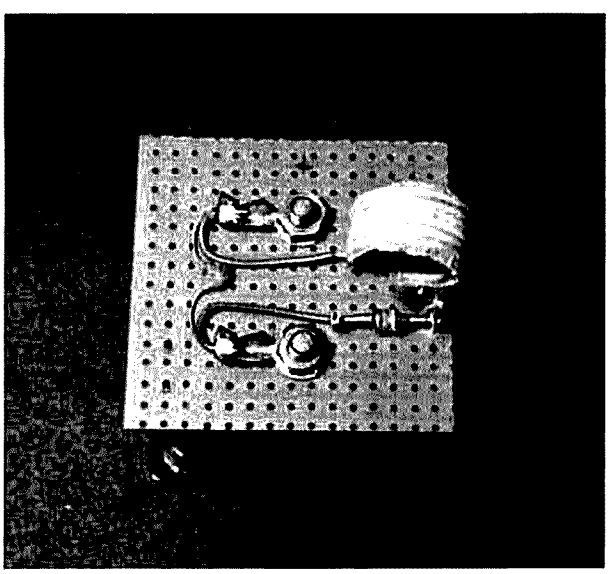


fig. 1. Circuit for the simple swr indicator. L1 consists of 20-inch piece of no. 14 double-cotton-covered wire closewound on 3/4-inch form.

The coil and diode were mounted on a piece of perf board, which was mounted directly onto the two machine screws at the rear of the meter. Watch the rectifier polarity – the positive end is connected to the positive side of the meter, and vice versa.

The only other component is a 1500-pF ceramic capacitor mounted across the meter terminals to bypass rf around the meter movement. The current is small, so solder well.

Business end of the simple swr indicator. An inductor, diode rectifier, and a few simple pieces of hardware make up the circuit.



operating principles

Quite heavy current flows in the center conductor of a coaxial cable. The same current also appears in the outside braid. The only difference is that the current is concentrated in the inner con-

table 1. Parts list for the inexpensive swr indicator.

quantity	Component
1	20-microampere meter
1	1N34A diode
1	.0015 μF ceramic capacitor
1	perf board, 2-3/4 in. square
2	meter connecting lugs, washers
3	.065-in. connectors
1 length	no. 14 dcc wire, 20 in. long polystyrene coil dope 1 x 2 in. plastic sheet solder and hook-up wire

ductor but is diffused to a great extent in the outer braid, which is capable of conducting about ten times the current in the inner conductor. However, the same amount of current is carried by the braid. It is this diffused, very minute current, that is indicated by the swr meter.

using the instrument

The outgoing current and any reflected current in the feedline will add and subtract in the feedline to form nodes (high current) or near-nulls (low current) at various points, which are generally unpredicatble. Sometimes a device called a transmatch is connected between transmitter and feedline to reduce the difference between these two currents; i.e., to tune the line. Such devices introduce some loss into the system.

The swr indicator described here is not an in-line device. It is slipped over the transmitter-end of the feedline, and the feedline reconnected to the transmitter. The transmitter is then turned on, but not keyed. Then, while the indicator coil is held in one spot on the line, the transmitter is keyed for length of time to make one dot. The meter should show an indication. The procedure continues as follows:

> Slip the device along the cable a few feet, and again key a dot on the transmitter. This reading will probably be higher than the first. Either way, it will indicate a standing wave on your feedline.
>
> Mark the feedline at one-foot intervals as far as possible with a light-colored crayon. Generally, there will be from 6 to 15 feet of line inside the shack that you can so mark. Move the coil along the line, key the rig and record the data. At one point you'll get a maximum reading, either side of which the indication will decrease. At another point on the line you'll get a minimum reading, either side of which the reading will increase. In my case, a node and null were obtained in just 6 feet of line, so I knew I had a high swr. If your swr is low, you may have to move the device along the line a greater distance to determine where node and null occur.

determining swr

Once you have recorded the readings on the meter, merely divide the lowest into the highest reading. The quotient is a sufficiently accurate indication of swr for most ham purposes.

My feedline runs outdoors parallel to the ground for about 20 feet before heading up to hang at a right angle to the dipole. I used a coat hanger bracket to hold the meter onto a wood pole; this arrangement was slid along the cable to take additional readings. I obtained the same node/null readings as on the indoor section of line, so this procedure may not always be necessary or convenient. These indications, of course, occurred at distances further apart at 40 meters than on the higher bands.

conclusion

If, by chance, your swr is *really* low; i.e., very close to unity, this fact would be immediately obvious due to the very slight difference in meter readings over a quarter-wavelength of line at the operating frequency.

The main advantage of this little device over an in-line instrument, such as the swr meter, is that no insertion loss or mismatch occurs between transmitter and load. Such loss may be significant at certain frequencies and power levels.

ham radio

fire protection

in the ham shack

With a few simple precautions the electrical and fire hazards in your shack can be minimized

Jack Darr and Alan James

Hot rocks and hot receivers are fine to have around the ham shack, but not when they are caused by a "hot house" – in other words, *fire*. Wherever there's electrical gear, there's always the chance of fire. Furthermore, there is always other fire-prone stuff either in use or lying around the busy shack.

It shouldn't take much talk to convince any serious ham he should make a fire-safety check at regular intervals. The only trouble with this plan is that many fire-prevention precautions are so darn simple that they get overlooked. Let's look over some of the commonest hazards, and work up a check list to keep from missing any.

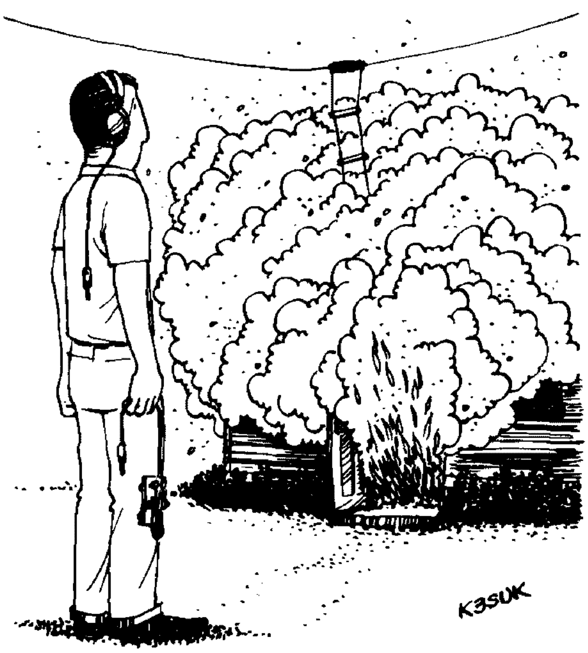
electrical fires

Hams have a special problem. A fire in the ham shack is always likely to be or become an electrical fire, because there is always electric power around. This situation demands special precautions. You can't just pick up a bucket of water and heave it on. (Assuming normal equipment in your shack includes a bucket of water – mildly unusual for a radio station, you must admit!)

First, what is an "electrical fire?" Well, it always begins with an arc. When in-

sulation breaks down, or the applied voltage is too high for the insulation, an arc "happens." Temperatures in even a tiny arc get pretty high – up into the thousands of degrees.

The heat can melt ordinary insulating



Fuse it . . .

material, crack ceramic insulators, and raise the dickens in general. It can even make insulators into conductors, by carbonizing them! Phenolic and similar insulators turn into carbon when they're burned, and away we go with more arcing. Carbon encourages the arc, which gets hotter, which carbonizes more of the insulator, and so on and on. The heat generated melts other insulation on nearby wiring – which can start still more arcs. You can wind up with a real mess.

You could build a chassis that is absolutely fire-safe, if you want to. Glass-insulated wire, ceramic insulators, chassis coated with Teflon, and liquid asbestos sprayed over the whole thing . . . all would help, but would skyrocket the cost and be a lot of trouble. There's an easier answer – just give the thing better protection. If you can stop an arc as soon as it starts, you're in; the unit will easily keep its own cool. One way is simply by removing the power.

fuses as friends

The easiest way to stop an electrical fire is to keep it from getting started. What that boils down to is fuse protection on *all* electrical equipment in the shack! You might have the dangerous idea that "Aw, what the heck! It's only 110 volts!" That little ol' 110 volts has all the current available it needs to burn your shack to the ground. Current is what makes an arc so hot and destructive. (An arc-welder puts very low voltage on its electrodes, but the dozens of amperes that flow in them can melt steel.)

Fuses are affected mainly by current. The voltage rating you see on a fuse is put there so you won't use it in some circuit where the voltage is so high that an arc could form *across the fuse* after the fusible wire inside has melted. The principle of correct fusing is: use a fuse that can carry the normal current load, but will open up instantly whenever there is a serious current overload (such as caused by an arc).

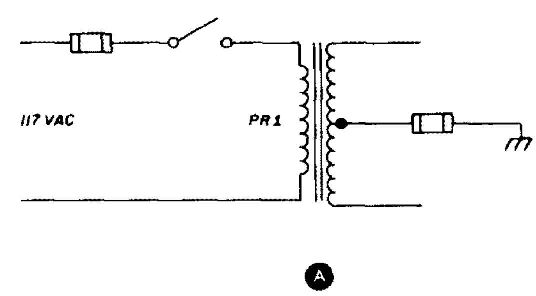
There are also small circuit breakers, which do the same job of protection. Chassis-mount circuit breakers come in ratings from 0.325-amp up to 7.0 amps. The kind that go in the ac line range from 5.0 amps to any size you need. However, the small ones will protect ham gear.

So, there's the first point for your fire-prevention check check list. Make sure every piece of gear in the place, down to the soldering iron, is protected by fuses of the right size! When you build

anything, even kits, if the designer didn't use fuses where he should have, you add them! There is a simply amazing number of units that have no protection at all, or not enough.

The diagram in **fig. 1A** shows two places where they'll do the most good. If your rig uses the popular bridge rectifier circuit as in **fig. 1B**, the secondary fuse is best put in the dc output circuit. A short in one of the rectifiers or filter capacitors would have to blow the primary fuse.

You need to know only one thing to add a fuse: how much current the equipment draws normally. Add a little safety factor of about 50%. Up to about 10 amps, the fuse rating can be about double the normal current; in the higher values, up to 15 or 20 amps, use a smaller margin.



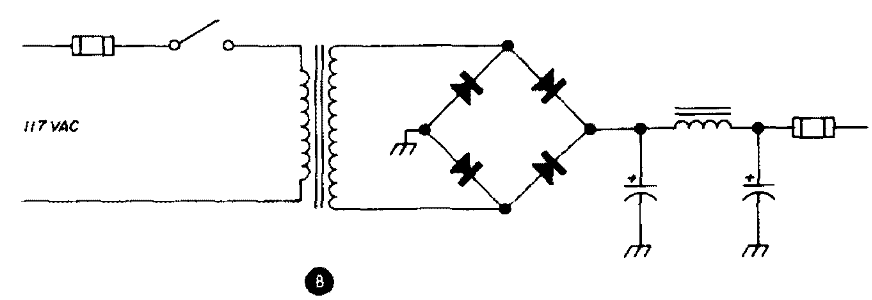


fig. 1. Methods of fusing power supplies.

To find out normal current for a B+ fuse, insert a dc milliammeter at the point where the fuse is to be connected. Primary fuses aren't that easy to figure, since ac ammeters aren't common. Easy way: hook a 1-ohm wirewound resistor (big watts) in series with the primary, and read the ac voltage across it with the equipment operating at full load. A 1-volt drop means 1 amp of alternating current is flowing (automatic Ohm's-law computer). On factory-built gear, there is a rating-plate, usually near the line cord. Divide the wattage by 115 to calculate the primary current.

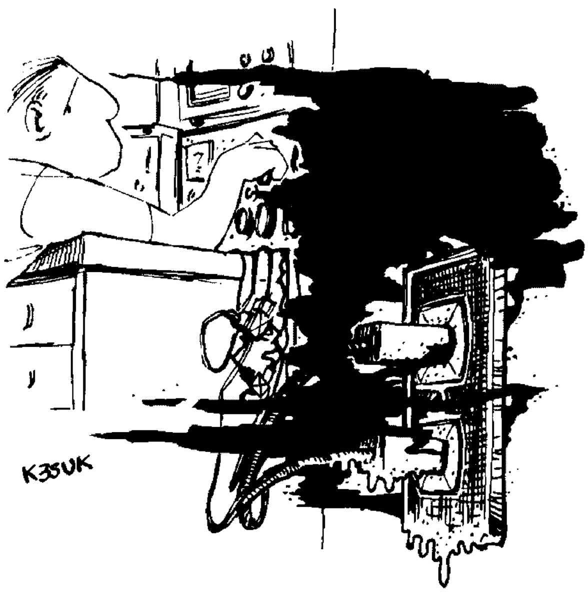
If you have some piece of gear that needs a fuse, but the chassis is so crowded it would be hard to mount, try a "fusible plug" on the line cord. A pair of glass cartridge fuses, which can be any size from ¼-amp up to 20-amp or more, fit into push-in clips. To replace a blown fuse, you just push it out of one end with a stick or kitchen match, and push the new fuse in.

So, the fuse is your answer. You can't usually clear a short once the arc has started, so you had better get power off that circuit as quickly as you can. The best way is with a fuse. Without current, the arc must stop. Result: no more fire hazard.

behind the wall outlets

An important factor in fire safety is the ac wiring. If you live in a house more than 20 years old, look it over. The fixtures (outlets, switches, etc.) used then weren't as good as those now. Besides, they're 20 years old. Wall outlets and

receptacles *were* rated at about 600 watts, which isn't much today. Plug in a big transmitter, a receiver, and a soldering iron, and you've had it.



Plug in a big transmitter, a receiver, and a soldering iron, and you've has it.

So what happens? The contacts (even if they're bright and clean, which they're not) get hot. This makes them tarnish, which makes them get hotter, which makes them get dirtier, which makes them . . . (sound of fire sirens approaching).

Modern outlets have a minimum rating of 15 amps, or 1800 watts, and you can buy 20-amp types for practically the same price. Contacts run cool.

The ac wiring in old houses is often only number 14 solid, which is not very heavy. Maximum current it can safely carry is a measly 4.6 amps. More than that makes the wire get hot. And there you have another fire hazard.

The shack can have surface-type fittings, with Romex cable along the surface of walls or floor, on the underside of the operating bench, etc. However, be sure the wire can carry your maximum load – and then some.

Don't start heavy shack-wiring at an existing outlet-box. This is worse than useless; it's dangerous. If you tap a wall outlet that happens to be at the end of about 50 feet of number 14 wire, you can heat up the old wire with the overload. If special ham-shack wiring is installed, go all the way to the main breaker-box or load-center, and run new heavy wire from there.

If you have a big rig, say a full-gallon transmitter plus a good-sized receiver, those babies are going to draw a pretty good hunk of current. With a kilowatt of rf coming out, the transmitter alone can use up 2000 watts. The receiver takes another 150 - 200. That's crowding even the best modern outlets.

It would be better to run a special line right into the transmitter cabinet. A pair of standard home-type circuit breakers (one in each side of the line) can mount on the back of the cabinet in a neat metal box. You can also use them as a master switch when you want to work on the transmitter.

the lightning hazard

Lightning is a fire danger you can't always do a whole lot about. With that nice, big four-element "lightning-rod" sticking up outside, you're apt to get a shot once in a while. The basis of all lightning protection is keeping the hot stuff outside the building and shuttling it off to ground as quickly as possible.

Thing number 1 is a good lightning arrester. You can't actually "arrest" lightning, but you can "steer it" some place where it will do the least damage, and where it won't set the place afire. An arrester carries the energy off to ground by providing an easy path for it. The connection from arrester to ground had better be short, heavy, and direct. A too-small ground wire will simply vaporize if it gets a solid hit.

The coaxial transmission lines so popular now are safer than older open-wire lines. You can make a good lightning arrester from the cable itself. Carefully slit the outer jacket, near where the line enters the building. Turn the insulation back, and wrap 8-10 turns of clean,

heavy, bare copper wire snugly around it. Don't solder it; you'll melt the plastic insulation. Cover this with at least two layers of tightly wrapped plastic tape, and spray the outside with a couple of coats of an acrylic plastic like "Krylon Clear." Run the other end of the copper wire *straight down* (don't make sharp bends in it) to an 8-foot ground rod, driven all the way in.

For open-wire transmission lines, a wide-spaced spdt knife switch can be used to disconnect the transmitter and ground the antenna when you're not around.

Even with these precautions, a direct hit can wander all over your incoming leads and wires. To minimize the chance of a fire starting, run *all* transmission lines well away from wooden walls, drapes, and any flammable materials. Even a coaxial transmission line can get very, very hot if it's carrying a lightning surge.

. . . an 8-foot ground rod driven all the way down.

(We've seen coax line welded to a 100-foot tower, all the way up!)

A system of good grounds is one of your best lightning-damage preventives inside the shack. Run a good, heavy wire from your transmitter cabinet (receiver, cabinet, too) over to a good ground – cold-water pipe if you must, but the deep rod outside is better. Upstairs shacks must use a very heavy ground cable; you can often get pieces of stranded pole-guy cable from the power company.

A surprising amount of energy is floating around on the ac power lines when you get smacked by lightning. So, proper-sized fuses can do a lot to protect your stuff. With a direct hit, the fuses will vaporize instantly (and the fuse-holder may, too – wait till you see a 2-pound ceramic fuse-holder blown to powder!). However, when they go, they'll leave the circuit open, and there won't be so much likelihood of arcing from the hot powerline to start fires.

diverse dangers

Watch the common things you're apt to forget. Take the soldering iron. It's handy and all that, but it does get plenty hot. Put yours on a heavy stand that can't tip over. A stand with a screen or mesh cover over the barrel is best. Keep rags or scratch-paper away from the barrel of the iron. Be a good housekeeper, and you'll be much more fire-safe.

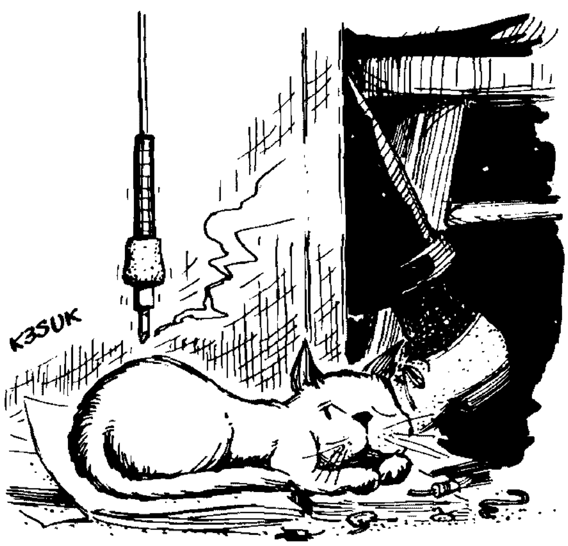
Now and then you have to use chemical compounds to clean things around the shack. Switches and such things get dirty, and you have to police-up so they'll work again. Be darn sure the cleaner is fireproof. Nobody in his right mind would use gasoline for this, but lighter fluid can be just as dangerous. A little vapor . . . a spark . . . and BOOM.

Carbon-tet is fireproof, but the vapor can make you pretty sick. The safe stuff is the spray-can cleaners. They're nonflammable, and there are no noxious vapors. Also, they do a better job of cleaning.

If your shack is in its own separate little building, a gas stove for heating can be dangerous, especially the open-flame type without a vent to the outside. Check the connecting hose at the gas outlet often, to make sure there's no leak. Natural gas is bad enough, but bottle-gas is worse in one respect: It's heavier than

air. If there's a leak, a layer of gas forms from the floor up. When the layer gets high enough, any spark (like turning on the shack lights) can set it off. Then, if you don't have a fire, you may have an air-conditioned shack – no windows, and maybe no door (and probably no roof).

Heating gas is odorized so you can smell a leak. So, if you think you smell a skunk when you walk in, keep your



**take the soldering iron . . .
it does get plenty hot.**

fingers off the switches until you find out what's causing the smell.

if prevention fails

We've mentioned the most likely conflagration-commencers, and how to keep them from starting a fire. What do you do if a fire suddenly blazes up anyway? Easy – grab your fire extinguisher and pour it on. That's fine, but . . . be darn sure you grab the right kind of fire extinguisher! The wrong kind can get you killed.

If there is any electrical hot stuff in or near the fire, and there probably will be somewhere, *DON'T* use a soda-acid, a foam, or a carbon-tetrachloride extinguisher.

The soda-acid types are those old ones you see hanging on the walls with a hose on top. You turn them upside down to spray the fire. The solution in them is mostly plain water. You can figure out what can happen when a half-inch stream of conductive water hits a live 2,000 volts, with you hanging onto the metal case. You'd better hope you're standing on dry rubber.

The foam types contain mostly water mixed with special chemicals. The popular carbon-tetrachloride types are okay if you're out-of-doors. Carbon-tet is a very good fire-extinguishing fluid, but when it strikes a hot fire, the fumes are deadly poison! They are phosgene, the toxic gas used during World War I. In a small room, and most ham shacks are small, it can put you off the air permanently in just a little while. So, don't use carbon-tet in the shack, unless you have time to stop and put on a gas mask!

There are three types of extinguishers recommended for use on electrical fires. The best known of these is the CO_2-type . . . or "fogbottle," as some people call them. (Don't confuse it with what a fireman calls a "fog nozzle" which is a special hose-nozzle that sprays plain water in a very fine mist.) The carbon-dioxide gas smothers a fire by keeping it from getting enough oxygen to burn. It is harmless to humans.

The second type is a dry-powder extinguisher, which uses plain bicarbonate of soda as the fire-snuffer. The powder is often packaged in long paper tubes; you tear off the cap and toss the powder at the fire. Better ones are pressurized, and spray the powder on the fire.

Freon gas, the propellent in most pressurized spray-cans, is used in another electric-safe extingusher. Freon provides its own pressure so is easy to spray at the base of the flame. It works just like carbon-tet, but does not liberate toxic fumes.

You can buy any of these three extinguishers in supermarkets and drug stores, in 1- and 2-pound capacities. These sizes are not recommended for big fires. The fire chief will tell you the best sizes are 5- and 10-pounders. These will

put out most fires completely, even though they have a good start.

fire alarms

If you're in the shack when a fire starts, you can combat it instantly. If something heats up while you're out of the shack, that's different. Some kind of automatic fire-alarm system can give you a warning that action is needed and needed right now!

The alarm itself can be a loud bell or anything else with an unmistakable sound. It can be ac-powered, or have a battery if you want a fail-safe system. You can rig up a little charger across the battery.

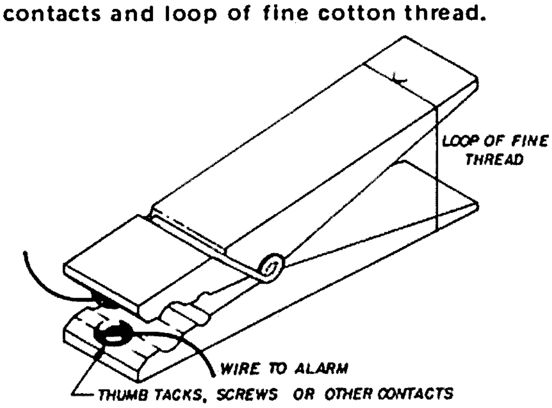
You'll need some kind of sensors that will detect heat, flame, or smoke. This can be done in dozens of different ways. You can have a lot of fun thinking up gadgets that will detect fire.

One of the simplest is shown in fig. 2. It's a clip-type clothespin swiped from the clothes-washing department. Put a couple of thumbtacks in the jaws for contacts. Then rig up something to hold it open until it senses fire. Years ago, a small loop of celluloid movie film was used, which burned instantly. Modern plastics aren't too good; most are fireproof. However, a loop of fine thread is okay. Use just enough to hold the jaws open against the spring tension. This thread will burn open almost instantly; test it to make sure.

Transistors are heat-sensitive, some amazingly so. When a transistor's temperature goes above a certain point, it will conduct heavily, often going into ava-

Heathkit GD-77 receiver alarm sounds high-pitched electronic tone when actuated by GD-97 transmitter.

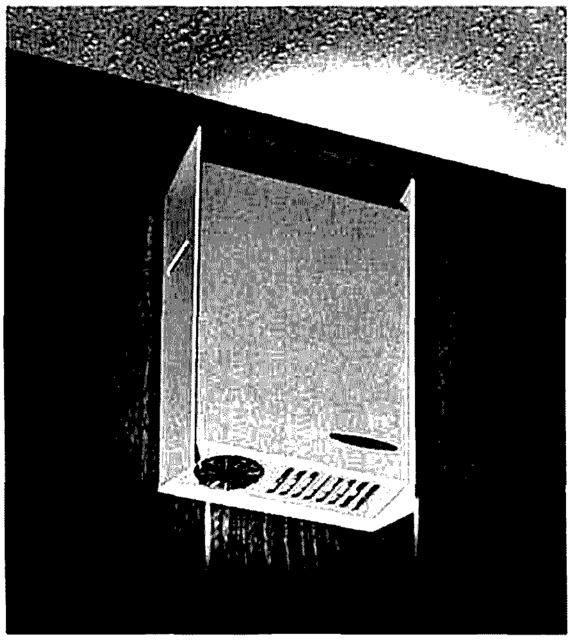
fig. 2. Simple heat sensor consists of household clothespin, thumb-tack contacts and loop of fine cotton thread.



FIRE-PROTECTION CHECK LIST

1. Electrical
   - (a) Does all equipment have fuses?
   - (b) Are they small enough to give real protection?
   - (c) Is all wiring heavy enough to carry its normal load without heating?
   - (d) Are all outlets and switches in good shape — not dirty or loose?
   - (e) Is the radio gear fully protected by suitable lightning arrester?
   - (f) Are your fire extinguishers the correct type for use on electrical fires?

2. Housekeeping
   - (a) Is the shack neat?
   - (b) Are papers, old magazines, or other flammable materials stacked around in corners?
   - (c) Worse still, are they stacked right on top of hot electrical equipment?
   - (d) Is the workbench soldering iron protected against rags or paper falling on the hot barrel?
   - (e) Is the wastebasket made of fireproof metal or of meltable plastic?
   - (f) Is the wastebasket full, running over, or kept nearly empty?
   - (g) How are the ashtrays? Full? Clean? Big enough for a long rag-chew?

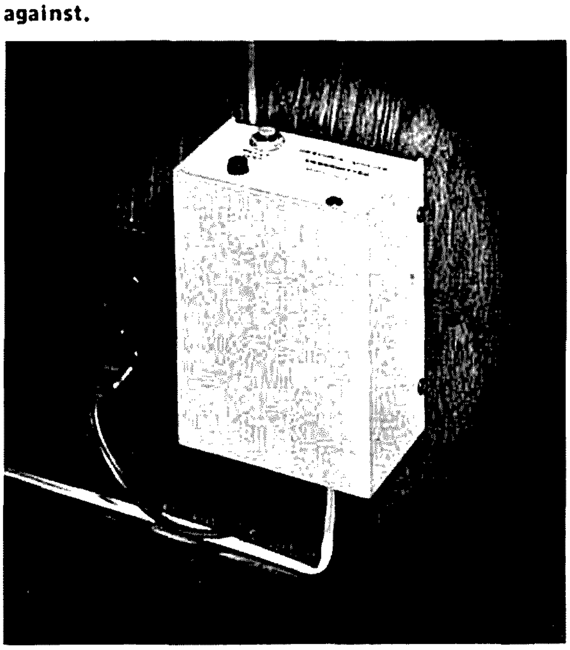


Heathkit GD-87 smoke detector actuates if smoke or 133° F heat is present.

lanche. You could connect the transistor across the input of a small dc-coupled amplifier. A change of resistance in the sensor transistor will trigger a stage that closes a relay.

Thermistors could be used, fastened to parts that might overheat. A bridge circuit will respond whenever the thermistor's resistance goes down as its temperature rises.

Heathkit GD-97 utility transmitter accepts any type switch or sensor to detect hazard you wish to guard against.



Smoke detectors can be simple: an amplifier circuit with a phototransistor at its input. Mount a pilot light to shine on it through a small light-tight tube. This setup is put just above any equipment where fire might start. If smoke comes up the tube, it will cut down the amount of light falling on the cell; this changes the cell resistance and . . . ding-a-ling-a-ling!

For maximum protection, make your system fail-safe. Design it to react, tripping the alarm, if its power supply or any of its parts fails. The principle is simple. Use a normally closed relay, and design the system so the relay is pulled in — energized — in the cool condition (no fire). If fire is detected, current is cut off and the relay drops out (closing the contacts). If the power-supply fails, or one of the amplifier parts should blow — same thing. No matter what happens, the relay armature is released, and the contacts close, sounding the alarm. The alarm bell must have its own separate battery supply, applied by the relay.

Ham ingenuity will bring out many other ways of doing this important job. Just look for some kind of device that responds to heat, or to smoke, and there you are. If you don't like to design things, there's a kit available from Heath Company; you can see it in the photo. It is wireless and has several kinds of sensors.

do it now

There are wise sayings we could finish with, like: "Don't throw lighted matches into the wastepaper baskets full of old message blanks." But this is kid stuff. If you're not smarter than that, you haven't read this far.

There is a surprising amount of unprotected gear around the average ham shack, if you get right down and look for it. When you find something like that, add the needed protection. It won't take long, it won't cost much, and the peace of mind will more than make up for it.

ham radio

mosfet converter for receiver instrumentation

The Heathkit SB-610 monitor scope can be used in the receive mode with this circuit

Robert A. Schiers, Jr., WA9ZMT, 5522 Laurel Hall Drive, Indianapolis, Indiana 46226

An old saying is that necessity is the mother of invention. In this case the adage is quite precise. I own a Drake TR-4 transceiver and wanted to use a Heathkit SB-610 monitor scope to observe signals in the i-f passband of the TR-4 receiver section. However, the TR-4 i-f is 9 MHz, while the vertical amplifier in the SB-610 is only capable of responding to signals up to 6 MHz. This article describes a simple local-oscillator/mixer I built to put the TR-4 receiver i-f passband where the SB-610 could handle the frequencies involved.

The SB-610 verticle amplifier has a peak sensitivity of 70 mV/inch (rms) of the vertical deflection if the vertical amplifier is tuned to 455 kHz. This is in contrast to 600 mV/inch at 6000 kHz. A mixer is obviously needed, and, as can be seen by comparing sensitivities, 455 kHz offers the best response.

A tube-type mixer involves cumbersome components for heater and plate power. I, therefore, decided on a transistor mixer, which requires only a small battery.

conversion gain

A conventional bipolar transistor mixer, operating in the driven-emitter configuration, requires substantial drive from the local oscillator. Furthermore, this circuit gives conversion gain. Conversion gain is defined as

$$\text{gain in dB} = 20 \log_{10} E_{out}/E_{in}$$

where

E_{out} = converted-frequency output voltage

E_{in} = amplitude of the voltage to be converted as presented to the mixer transistor base

After reading some articles on the dual insulated-gate mosfet mixers, I decided that this device fulfilled the goals of this project; i. e., simplicity, light loading of the local oscillator by the mixer, and a conversion gain of up to 18 dB.

the circuit

The converter schematic is shown in **fig. 1**. The LO signal is applied to gate 2 of the 3N159, which must be biased to 4 volts by the divider across this part of the circuit. Both gates of an igfet present a very high impedance to the driving signal, usually of the order of 100k ohms. The rf signal to be converted is applied to gate 1, where no special biasing is needed. However, this gate is tied to ground through a high resistance to protect the device from static charges or transient voltages.

The 3N159 drain contains two 455-kHz i-f cans. The first i-f can is in the drain to present a high impedance to the converted signal. The other i-f can is used to match the low-impedance output of the first i-f can to the high-impedance input of the tuned vertical amplifier of the SB-610.

* The LO may be obtained from International Crystal Manufacturing Co., 10 North Lee, Oklahoma City, Oklahoma 73102. The type OX in kit form is about $3.00. Crystals for any frequency can be ordered for about $4.00. The 3N159 is available from Allied Electronics, 100 North Western Ave., Chicago, Ill. 60680. Cost is about $2.00 in small quantities.

local oscillator

The diagram shows a typical circuit available from International Crystal.* The oscillator is specified for 0.2 volt across a 50-ohm load. The obvious impedance mismatch is unimportant in this application, because distortion and power transfer are not an issue when trying to obtain a voltage swing across such a high impedance as that presented by the igfet gate.

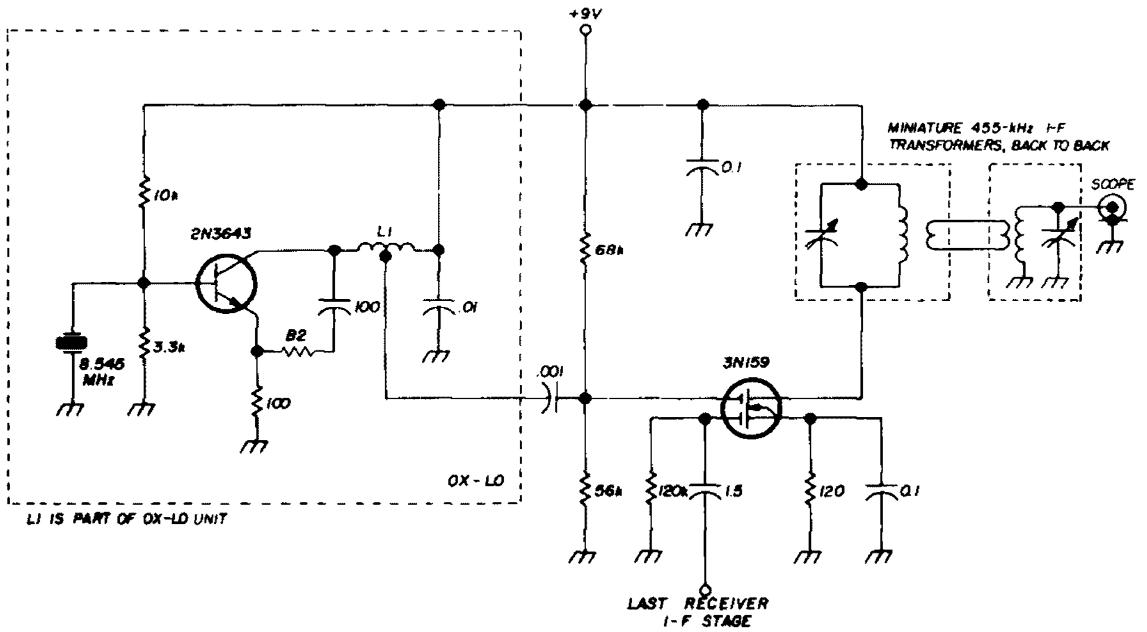


fig. 1. Schematic of the 9-MHz to 455-kHz i-f converter for driving a Heathkit SB-610 monitor scope in the receive mode. The igfet mixer provides 18 dB conversion gain with very low injection voltage.

construction and operation

Construction is not at all critical, nor is the supply voltage. The conversion frequency can be changed easily by tracking the LO frequency 455 kHz below the incoming rf signal. I varied the voltage applied to both the igfet and the oscillator from 4 to 9 volts without any degradation of signal quality or amplitude. Current drain varied from 12 mA at 4 volts to 40 mA at 9 volts. The voltage for the 3N159 can be varied anywhere between 5 and 9 volts with no noticeable deterioration in output amplitude or quality.

The current drain for the entire circuit is approximately 40 mA, depending on

supply voltage. Despite the impedance mismatch, no problems were encountered in driving the very high impedance of the 3N159 gate.

The coupling capacitor from the plate of the last i-f stage in the receiver was selected for the smallest value necessary to ensure adequate display height on the SB-610.

The reason for the dual 455-kHz i-f transformers was based on impedance-transformation considerations. The minature transformers available to me had a high-impedance input with a low-impedance output, whereas the input of the SB-610 is about 100k ohms.

precautions

The igfet comes with a shorting clip that connects all leads together. This clip should be left on the device until it is placed in the circuit. Don't allow your fingers to come in contact with any part of the device except the case during installation. A small piece of solder can be wrapped around all leads, then the solder can be removed after installation.

final observations

The RCA 40673 dual-insulated-gate mosfet is zener-protected and may be directly substituted for the 3N159 *in this circuit.* Such substitution is highly recommended, as the special-handling problem of the device is eliminated.

If at any time you think you may have damaged the 3N159, the device can be easily checked. With a drain voltage of 10 volts and 120 ohms in the source; and with both gates tied to ground through a high resistance (100k), a drop of a few tenths of a volt should be observed across the source resistor. In addition, if gate 2 is biased up 1 or 2 volts, the current will rise sharply to 4 mA or greater, depending on source resistance and drain voltage. Note that the characteristics described pertain to *n-channel depletion-mode fets* only!

I'd like to express my thanks to Leo Bovard, W9SCK, for his help in the development of this project.

ham radio

the ham notebook

testing unknown meters

A typical 10-megohm or higher vtvm is an excellent tool for testing unknown meters to determine their approximate current ranges without the danger of pinning the needles or burning out the windings.

Set the vtvm to OHMS and the range dial to R x 1 MEG. Connect the test leads to the unknown meter. Move the range dial, one step at a time, through R x 100k, R x 10k, etc., until the meter under test reads upscale.

The maximum current through the meter from a typical 10-megohm vtvm (Lafayette KT-174, for example) is as follows:

range	I max	vtvm 1% series res.
R x 1 meg	0.156 μA	10 meg
100k	1.56 μA	1 meg
10k	15.6 μA	100k
1k	156.0 μA	10k
100	1.56 mA	1k
10	15.6 mA	99.1*
1	170.0 mA	9.1*

*0.9 ohm is allowed for the internal resistance of the D cell and test lead.

The effect of the meter resistance on the first four ranges of the vtvm is usually negligible. Don't try this idea with a vom. The currents will almost surely be much greater than those shown above, and the meter being tested may be destroyed.

Harold E. Brown, W1ONC

adjustment screwdriver

Many transceivers have internal adjustments for the S-meter, final-amplifier bias, carrier null, etc. Trying to get a screwdriver into these adjustments through the perforations in the cover is next to impossible. Here's what I use.

I fashioned an ordinary wire coat hanger into an adjustment tool by clipping the hanger about 1½ inches each side of where the Y is formed. Then I straightened this section until it was perpendicular to the hook. Next, I straightened the hook and filed the end until it was flat enough to fit into the transceiver adjustment slots. The finished tool should have the appearance of a T.

The tool is small enough and long enough to go through the holes in the transceiver cover, and it is strong enough to withstand the torque of some of the more stubborn adjustments. It's also handy for tapping those sticky change-over relays that hang up once in awhile.

Dan F. Davis, WAØKGS

75A-4 modifications

To anyone who has owned and cherished a 75A-4 receiver for many years, as I have, the discovery that it has developed sensitivity and frequency-stability problems is like discovering a trusted friend to be unfaithful. Here are some solutions to these problems, plus a hint to improve the receiver's audio response.

insensitivity

Loss in sensitivity first appeared as a loss of one or two S-units after about an hour of operation. (I use the "calibrate" signal as a sensitivity reference for a specific S-meter reading at 14.2 MHz, with a 50-ohm dummy antenna.) Sensitivity loss gradually increased until it was 6 or 7 S-units after 15 minutes of operation.

After several frustrating weeks of signal tracing and a new set of tubes, I was about to give up when I stumbled onto the answer. The "rejection tuning," which is a bridged-T filter, has a sharp, deep null when properly adjusted. I noticed that when the set was first turned on, the "rejection tuning" behaved normally, but after warmup the null deteriorated and finally became useless. It seemed as though a comparitively low resistance was across the bridged-T inductor, L26.

The schematic shows one-half of V7, the Q-multiplier tube; some resistors; a choke; and C71, a 1000-pF capacitor in series across this inductor. Checking with a vtvm between ground and either side of L26 showed a *positive* voltage, which varied between 2 and 5 volts. I disconnected C71 from the inductor, but left the other end connected to the plate of V7 (and hence, B+). When I touched the vtvm probe to the free end of C71, I found the positive voltage to be even higher and still varying.

I checked other grids in the receiver with the vtvm and found four other leaky capacitors. When these were replaced, the sensitivity problem cleared up completely. A vtvm must be used, however. A 20k-ohms-per-volt vtvm simply shorts the leaky voltage to ground and gives no indication.

probable cause

When checking for leaky capacitors the set must, of course, be on and the rf gain control positioned fully clockwise. If the rf gain control is backed off, the higher bias voltage will swamp the leakage voltage on some of the grids. I can only conjecture as to the cause of the capacitor leakage; but my friend Ray Wood, W9SDY, suggests migration of the silver coating on the mica. Until a better explanation comes along, I'll accept Ray's theory. Migration is supposed to accelerate in the presence of a dc voltage. However, some unused mica capacitors of the same type and vintage showed leakage to some degree; new units showed no leakage at all.

frequency instability

The frequency instability problem became evident in exactly the opposite manner. The pto frequency jumped around for about ten or fifteen minutes after the set was turned on, then settled to its usual rock-steadiness.

According to an article in *QST*,[1] instability in the pto can be attributed to several factors, including capacitor C205, a 51-pF mica of the same type as those giving sensitivity problems. When I originally read the *QST* article, I assumed C205 might be changing value. However, after the previous experience with leakage, I found that a vtvm check between pin 1 of V15 and ground showed the telltale positive voltage.

After the set had been on for 10-15 minutes, the positive voltage disappeared, and the pto became stable. I suspect that, since a current flows through the capacitor when the pto is operating, the leakage path burns off after several minutes, then regenerates when the set is turned off. At

any rate, the pto remained stable after installing a new capacitor for C205.

audio response

I became increasingly annoyed by an emphasis on the bass of the audio response as both the 75A-4 and I grew older. I realize that as we grow older our ears become less sensitive to the higher frequencies, but retain their response to the lower frequencies, thus accenting the bass.

A newly acquired audio generator provided the opportunity to check the 75A-4 audio response. A surprisingly high peak (10 dB) appeared at 100 Hz, which decreased sharply on either side, to a level output between 300-3000 Hz.

Reference 1 suggested that feedback resistor R71 be removed. I tried this when I first read the article, but I didn't like the increase in audio, which required riding the af gain control; nor did I like the unpleasant audio quality.

The audio-signal generator showed that, with resistor R71 disconnected, the 100-Hz peak disappeared and the entire audio response became a broad peak centered around 3200 Hz. Substituting a 100-k resistor for the 33k originally used for R71 eliminated the objectionable 100-Hz peak and smoothed the entire audio response. A 1-meg resistor in series with the af gain control improved its action.

less bass response

Another source of bass emphasis is the cathode network associated with the noise-limiter diode, V 12. Since the limiter is useless for cw and ssb, I removed V 12 and inserted the leads of a 0.005-μF capacitor into the tube socket (holes for pins 2 and 7). This connects the detector output directly to the af gain control and eliminates the cathode network. If you would like even less bass, use a 0.001 μF capacitor.

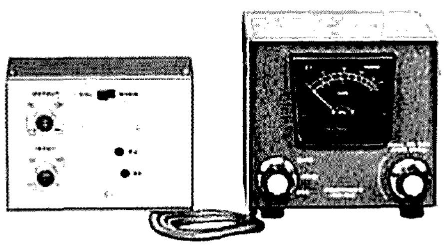
reference

1. P. Rockwell, W3AFM, "Station Design for DX," *QST*, November, 1966, p. 53.

Albert G. Shafer, W4SD

new products

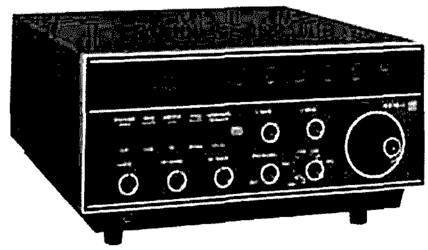
heathkit wattmeter/ swr bridge



The new Heathkit HM-102 wattmeter/swr bridge covers the frequency range from 1.8 to 30 MHz with wattmeter accuracy ±10% of full-scale reading*. The power capability of this new instrument is 2000 watts with full-scale readings of either 200 or 2000 watts. The built-in swr capability allows proper antenna tuning and transmission line impedance matching. The unit has a nominal 50-ohm impedance and provides negligible loss when inserted in a 50-ohm line. The remote torodial-type detector permits placement of the meter in a convenient location. The HM-102 kit is priced at $29.95 from the Heath Company, Benton Harbor, Michigan 49022, or use *check-off* on page 94.

*When calibrated on the 40-meter band through 300 feet of RG-58/U into a 50-ohm load.

collins receiver



The new Collins 651S-1 is one of the most versatile general purpose receivers ever developed, providing highly reliable reception from 0.4 to 30 MHz, with tuning phase-locked in 100-Hz steps. Because of the unique tuning system of the 651S-1, the convenience and "feel" of traditional receiver tuning is not sacrificed as a result of using discrete 100-Hz frequency steps. Large changes in frequency are aided by the 0.1-MHz and 1-MHz selector knobs.

Capabilities of the radio include computer programming, independent sideband, narrow band fm, operation from a remote control head, compatability with secure voice modems, phase-locked or manually adjusted vbfo, and syllabic squelch providing better reception of a weak voice signal in the presence of background noise.

Frequency readout on the front panel is provided by six light-bar devices, each of which consists of a seven-bar fluorescent segment tube.

The receiver construction techniques include Collins multilayer circuit boards mounted in plug-in card installations with blue-line connections. Micro-electonics logic and linear circuits are utilized. The plug-in card technique, along with convenient arrangement of other parts of the radio, provides easy accessibility. The receiver has no mechanical linkages – all tuning and band switching is performed electronically.

For information on the new 651S-1, write to Collins Radio Company, Cedar Rapids, Iowa 52406, or use *check-off* on page 94.

slow-scan television equipment

Robot Research, Inc., has announced a system of slow-scan television equipment that enables you to transmit and receive live television pictures with your existing station equipment. The model 80 slow-scan camera uses an efficient sampling technique that permits operation with everyday room lighting and transmits clear sharp pictures without involved adjustment. Features include digital timing chain, vidicon camera tube, built-in modulation calibration and solid-state construction.

The model 70 slow-scan television monitor demodulates and displays pictures transmitted by another amateur radio station. Simple connections between the monitor and your receiver are all that is required to receive an sstv picture. Included with the monitor are all of the switches and interconnections required to integrate sstv into your station. The monitor features balanced discriminator, automatic sync threshold, phase-locked horizontal sweep and solid-state construction.

The model 80 slow-scan camera and model 70 slow-scan monitor are available from Robot Research, Inc., 1250 Prospect Street, La Jolla, California 92037, or use *check-off* on page 94.

cubical-quad book

Bill Orr's new book, "Cubical Quads," features a wealth of new information not available in the previous edition. The new 2nd edition includes revised quad gain figures, analysis of the quad vs yagi, 4- and 5-element monster quads, miniature quad construction and performance, improved tri-gamma match for three-band quads, charts with dimensions for single- and multiband quads for 6 through 80 meters, as well as construction details for the delta quad, Swiss quad and birdcage quad.

This new book also includes information on feeding your quad for maximum swr; clear details on matching procedure; use of baluns; hints for sturdy, inexpensive quad construction; and how to adjust your quad for strongest signals. Quad installation, adjustment, evaluation and maintenance is covered along with quad facts; angle of radiation, directivity, power gain and front-to-back ratio.

''Cubical Quads'' is $3.95 from Comtec, Book Division, Box 592, Amherst, New Hampshire 03031.

heathkit frequency counter

The new Heathkit IB-101 digital frequency counter provides accurate counting from 1 Hz to over 15 MHz with five-digit readout. The Hz/kHz selector switch and overrange indicator give eight-digit capability. The new counter features high input impedance, automatic trigger level for with range input without adjustment, storage circuitry for non-blinking, no-count-up adjustment, and temperature compensated crystal-controlled time-base oscillator. The *automatic decimal locator* puts the decimal point in the right place for any measurement with no need for interpolation or figuring.

The IB-101 uses 26 integrated circuits, 7 transistors, 6 diodes and one mosfet. The dual-gate diode-protected mosfet in the input circuit provides proper triggering over a wide range of input levels. It will operate properly with input levels

from less than 100 mV to greater than 200 volts (depending on frequency). There is no input adjustment, and the instrument cannot be damaged by inputs within the specified range. Input impedance is 1 megohm shunted by less than 20 pF.

The new counter goes from kit to finished counter in about five hours, and complete adjustment can be made in a few minutes using only an a-m radio. No exotic test equipment is necessary. The IB-101 kit is complete with input cable. Priced at $199.95 from the Heath Company, Benton Harbor, Michigan 49022, or use *check-off* on page 94.

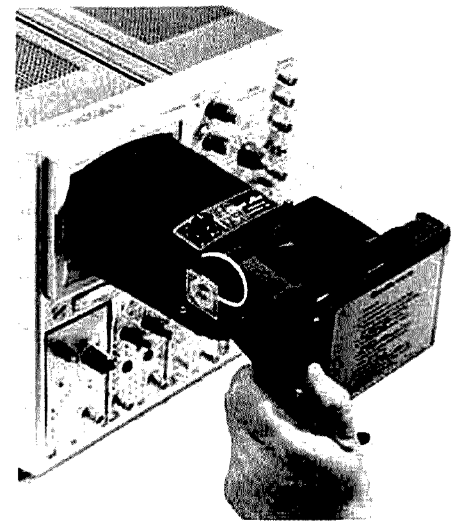
wideband rf power amplifier

The RF Communications has announced a new RF-806 linear amplifier module that features flat response within ±1 dB from 50 kHz to 80 MHz, and 3-dB bandwidth of 100 MHz. Gain is 47 dB with class-A amplification providing 10 watts with low harmonic and intermodulation distortion.

Input and output are overload protected against overdrive or operation into a short or open circuit. Unit is all solid state and operates from +28 Vdc. Standard options include 75-ohm input impedance and extended frequency response to above 100 MHz or down to 14 kHz. Accessories include impedance-matching transformers and power combiners.

Applications include use as a subassembly in test systems or amplifier chains to raise the power level of signal sources and generators without tuning or bandswitching. A-m, fm, pulse, ssb and other modulated signals can be amplified to 10 watts PEP with minimum distortion. Price is $795. For more information, write to RF Communications, Inc., 1680 University Avenue, Rochester, New York 14610, or use *Check-off* on page 94.

scope camera



Integrated Controls, Inc., has announced a low cost, hand-held oscilloscope camera set, Scope-Mate, which fits virtually all oscilloscopes. Fitting either three, four or five-inch round or rectangular scope faces, Scope-Mate can capture and record – either one shot or recurring trace – oscilloscope data. A standard Polaroid Colorpack II or III camera, in conjunction with the Scope-Mate hood, provides high-quality, high-contrast, black-and-white oscilloscope pictures in 15 seconds. The object-to-image ratio is one to one.

The Scope-Mate electronic control provides automatic adjustment of camera shutter speed, allowing you to vary photo contrast for perfect pictures.

The camera set, including a Colorpack II camera and Scope-Mate hood sells for $119.50. The Scope-Mate hood is available separately for $89.50. Order from Integrated Controls, Inc., Post Office Box 17296, San Diego, California 92117, or use *check-off* on page 94.

darlington power transistors

Five new series of complementary, power Darlington transistors – the first available in the industry – have been introduced by Motorola. The new silicon

power transistors provide gain up to 2500 (typical) and are available in current ratings from 4 to 16 amperes. With their high gain, the power Darlington transistors can be driven with integrated-circuit milliampere output current levels. The new devices are supplied in 60 and 80 V collector-emitter breakdown voltage ratings (a 100 V rating is available in the 16 A series).

The Darlingtons incorporate both driver and output transistors plus necessary resistors in one monolithic structure. Thus, the usual separate driver and associated emitter-base resistors employed in a Darlington amplifier are eliminated.

Darlington transistors are useful in relay and solenoid drivers, servo amplifier, audio amplifier and power supply regulator circuits. Because they're available in both polarities, they can be used in either positive or negative ground systems. And, used in pairs, they are especially useful in complementary applications such as direct-coupled pairs in audio amplifiers.

4 ampere MJ4000/MJ4010 series, 1000 minimum gain at 1.5 amp, 60 or 80 BV_{CEO}, 75 watts, TO-3 package.

5 ampere MJ900/MJ1000 series, 1000 minimum gain at 3 amps, 60 or 80 BV_{CEO}, 90 watts, TO-3 package.

5 ampere MJE1090/MJE1100 series, 750 minimum gain at 3 or 4 amps, 60 or 80 BV_{CEO}, 70 watts, plastic thermopad-type package.

10 ampere MJ2500/MJ3000 series, 1000 minimum gain at 5 amps, 60 or 80 BV_{CEO}, 150 watts, TO-3 package.

16 ampere MJ4030/MJ4033 series, 1000 minimum gain at 10 amps, 60, 80 or 100 BV_{CEO}, 150 watts, TO-3 package.

For more information, use *check-off* on page 94, or write to Technical Information Center, Motorola Semiconductor Products, Inc., Box 20924, Phoenix, Arizona 85036.

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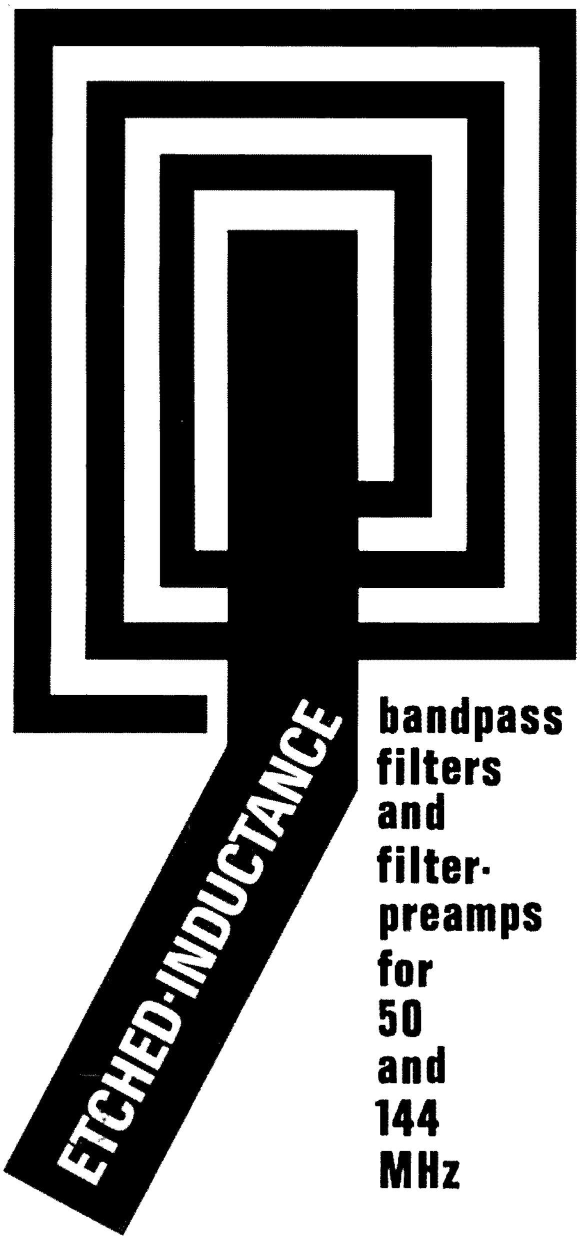

bandpass filters and filter-preamps for 50 and 144 MHz

this month

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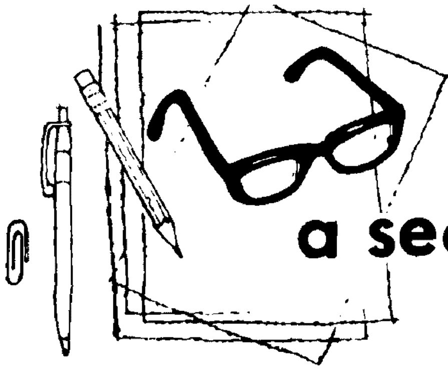


a second look

by jim fisk

As I **mentioned in this column** several months ago, microminiaturization is forcing inductors out of electronic circuitry. The sad fact of the matter is that it is virtually impossible to compress the inductor. Current inductance designs are much too large to fit into the rest of the circuit, and if the inductor is shrunk too much, performance rapidly deteriorates. Several ingenius substitutes have turned up to replace this bulky component: surface-wave devices (discussed in the September, 1970 issue of *ham radio*), active filters, integrated-circuit phase-locked loops, capacitor-loaded gyrators and etched-circuit inductors.

etched inductors

Although the etched-circuit inductor is still much too large for many applications, it results in considerable space saving, and unlike some of the other inductor substitutes, etched inductors can be easily built in the amateur workshop. One excellent application of etched inductors is discussed by W5KHT in his article on 6- and 2-meter bandpass filters and preamplifiers in this issue. Since writing the article, Bob has come up with a similar design for 220 MHz, as well as a completely etched-circuit two-meter converter that includes printed-circuit inductors. If there is sufficient reader interest in the etched-inductor two-meter converter, we will publish full details in an early issue.

Although printed-circuit spiral-shaped inductors are not new, they have seen limited use, even in commercial and military equipment. However, now that the ice has been broken, I suspect we will be seeing more and more of these inductors in amateur-built equipment.

The inductance of an etched inductor depends upon surface area of the spiral, conductor width, length of the spiral and number of turns. Since it's a rather complex calculation, the best bet is cut and try. One of the best materials for experimenting with etched inductors is the adhesive-backed copper-foil strip available from Cir-Kit. This material can be easily arranged into the square spiral, increasing the number of turns to increase inductance. Once you have arrived at the proper number of turns for your application, you may translate the design to a printed-circuit board if you want.

other techniques

The gyrator is a directional phase changer in which phase changes in opposite directions differ by 180 degrees. When loaded with a capacitor it has all the electrical characteristics of an inductor. Present integrated-circuit gyrators simulate inductance over the frequency range from dc to 20 kHz with stable Qs up to 1800. However, experimental designs at Bell Labs have provided adequate temperature stability and Q up to 100 MHz.

More work needs to be done before practical high-frequency gyrators are a reality, but research has been slowed by the high success of active filters that use low-cost ic operational amplifiers. Integrated-circuit phase-locked loops are also being used for miniature resonant circuits, as are surface-wave devices and micro-crystal filters. At this point in time it is difficult to guess which technique will provide the tuned circuits for the miniature communications equipment of the future, but lab work in the next decade probably toll the end of the inductor as we know it today.

Jim Fisk, W1DTY
editor

etched-inductance bandpass filters and filter-preamplifiers for 50 and 144 MHz

Easy to build, high-performance, narrow-bandwidth interdigital filters and bandpass preamplifiers that feature etched-circuit inductors

Bob Cooper, W5KHT

Although etched-circuit inductances have been around for over 40 years, very little has appeared in amateur radio publications. In 1929, Charles Ryder, an Australian, was granted a patent[1] that covered the earliest known application of an inductance, "applied by printing, gold blocking, painting, metal spraying or electro-deposition to the surface of a dielectric base."

In a sense, the Ryder invention was the forerunner of all modern-day printed-circuit devices, but more particularly, the thin-film etched inductances used in microwave integrated circuits. As a mass-production technique. etched inductances are about the neatest thing to come down the electronics pike since the transistor.

In the August, 1970 issue of *ham radio* I described an interdigital preamplifier[2] that married a grounded-gate fet preamplifier to a pair of comb-line filters. The resulting system produced a narrow-passband preamplifier that reduced overload and cross-modulation problems in high performance low-noise vhf converters.

After putting together a number of these units, it became apparent that although the completed devices were electrically adequate, any mass production for commercial use was unlikely because too much hand labor was required in their construction and alignment. The search was on for a mass-production technique that would provide the same performance as the costly hand-built strip-line unit. The etched-circuit Interdigital Series Bandpass Filters and Interdigital Preamplifiers described in this article are a result of that search.

bandpass filter circuit

The circuit shown in **fig. 1** was developed to satisfy the requirement for a bandpass filter which could be placed ahead of one or more grounded-gate fet amplifiers. This circuit* is a variation of the three-element T-section bandpass filter. The circuit was developed specifically for the etched-circuit inductance system, and interestingly enough, it almost is impossible to make it work without etched inductors.

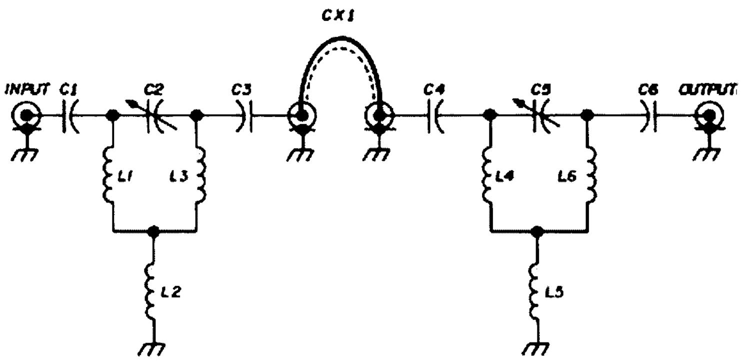


fig. 1. Circuit diagram of the interdigital series bandpass filter. Component values for 50 and 144 MHz are given in tables 1 and 2.

table 1. Parts list for the 50-MHz interdigital bandpass filters.

C1, C3, C4, C6 6.0 pF, 5% disc or tubular ceramic capacitors

C2, C5 4-40 pF midget trimmers (Elmenco-Arco type 422)

L1, L3, L4, L6 etched inductances on printed-circuit board (fig. 3)

L2, L5 15 turns no. 16 solid copper, close-wound on 1/4-inch form, with last two turns (ground side) separated from balance of coil by one turn width

CX 2½ inches RG-58/U coaxial cable, with braid twisted into pigtails so that total length of inner dielectric is 1-3/4 inches; pigtails are 1/2-inch long

table 2. Parts list for the 144-MHz interdigital bandpass filters.

C1, C3, C4, C6 2.2 pF, 5% disc or tubular ceramic capacitors

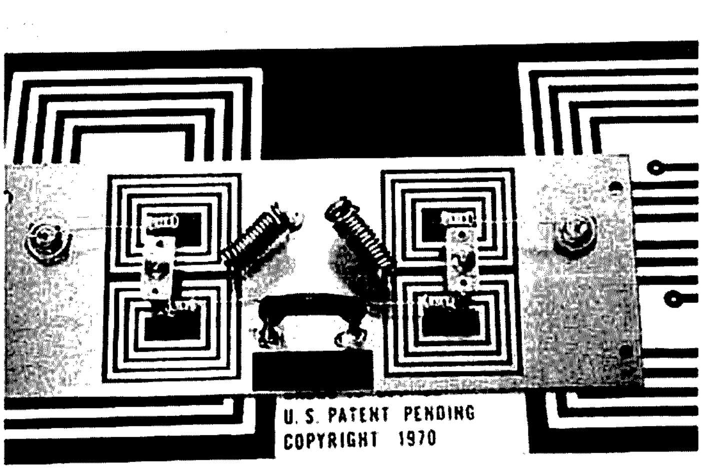
C2, C5 4-40 pF midget trimmers (Elmenco-Arco type 422)

L1, L3, L4, L6 etched inductances on printed-circuit board (fig. 4)

L2, L5 5 turns no. 16 solid copper, close-wound on 1/4-inch form with 0.2 inch standoff pigtails

CX 2-1/2 inches RG-58/U coaxial cable, with braid twisted into pigtails so that total length of inner dielectric is 1-3/4 inches long; pigtails are 1/2-inch long.

50-MHz etched-inductance bandpass filter.



In the circuit shown in **fig. 1.** two mirror-image three-section capacitively-tuned T-section filters are cascaded through a short length of coaxial cable. By careful selection of L3, L6, C1, C3, C4 and C6, the bandpass window can be

*The information presented here covers relatively narrow bandpass devices suitable for amateur applications. Etched-inductance interdigital bandpass filters and etched-inductance interdigital preamplifiers are the subject of patent applications filed by the author. Amateur construction of the units shown here for personal use will not violate the validity claims of the pending patents. editor.

varied from as narrow as 250 kHz at 50 MHz to as wide as 10 MHz at 200 MHz (or as narrow as 1.0 MHz at 200 MHz to as wide as 10 MHz at 50 MHz). Insertion been selected, total alignment time with a sweep generator is about 5 minutes. If you don't have a sweep generator and marker oscillator, the unit can be aligned

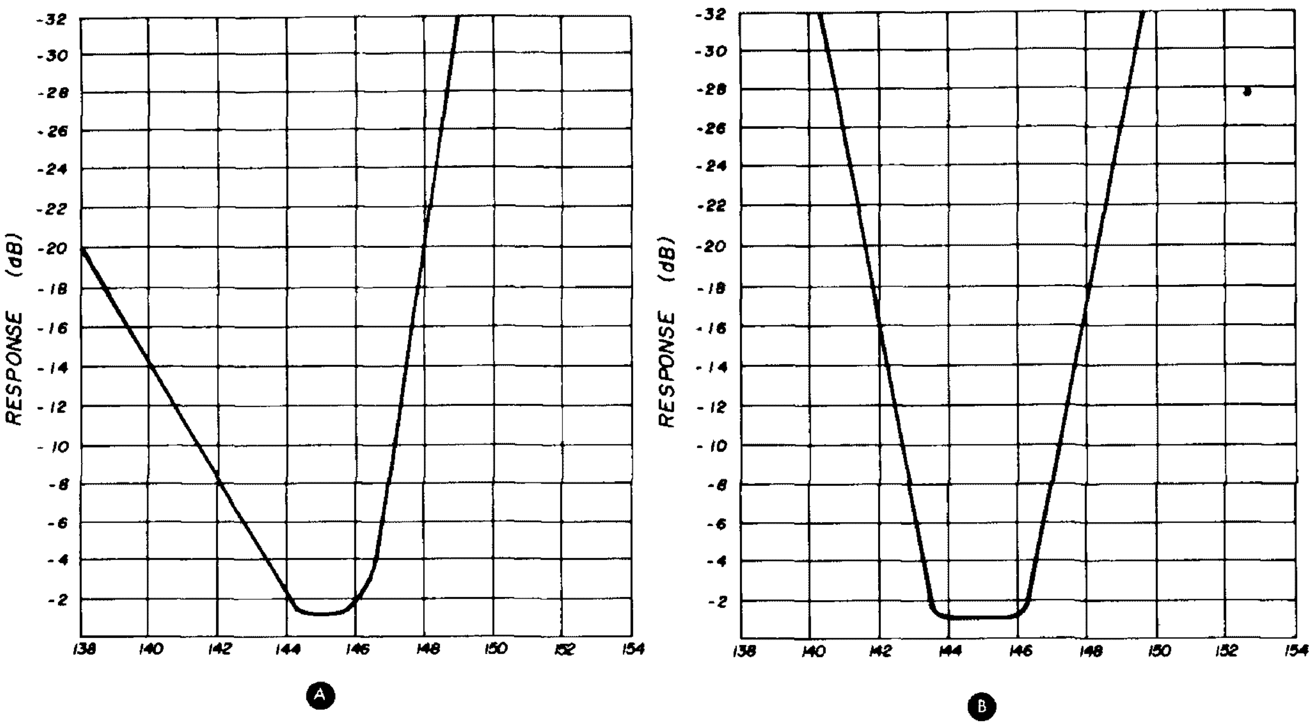


fig. 2. Bandpass characteristics of the 144-MHz etched-inductance bandpass filter. Filter may be adjusted for steep skirts on one side (A) or both sides (B).

losses are as low as 0.75 dB at either frequency.

In addition, the filter can be tuned so that is has very steep skirts on just one side (fig. 2A), or steep skirts on both sides (fig. 2B). Once the components have

table 3. Component values for narrow-band versions of the bandpass filters.

50 MHz	
C1, C3, C4, C6	4.5 pF, 5% disc or tubular ceramic capacitors
L2, L5	16 turns no. 16 solid copper, closewound on 1/4-inch form, separating two turns from balance of coil by one turn width
144 MHz	
C1, C3, C4, C6	1.8 pF, 5% disc or tubular ceramic capacitors
L2, L5	6 turns no. 16 solid copper, closewound on 1/4-inch form with 0.2-inch standoff pigtails

with nothing more exotic than a signal source and a receiver S-meter – in just about the same amount of time.

Etched-inductance interdigital bandpass filters for 50 and 144 MHz are shown in the photographs. A half-size layout of the 50-MHz board is shown in fig. 3; a half-size layout of the 144-MHz board is shown in fig. 4. Parts lists for the two boards are given in tables 1 and 2, respectively, for the 50- and 144-MHz filters. Commercially-made printed-circuit boards and complete parts kits are available.*

The 50-MHz filter has a 3-dB bandwidth of 1.0 MHz, centered on 50.250 MHz. The insertion loss between 49.8 and 50.7 MHz is 1.0 dB or less; filter rejection at 55.25 MHz (channel 2 video carrier) is as high as 40 dB. This should provide adequate front-end protection under the most taxing cases of channel-2 interference. If not, two filters may be cas-

caded together for up to 60-dB suppression of the 55.25-MHz signal. Insertion loss with two cascaded filters is approximately 2 dB.

MHz away from the center frequency will be on the order of 30 to 40 dB. A list of parts for the narrow-passband models of these filters is given in **table 3.**

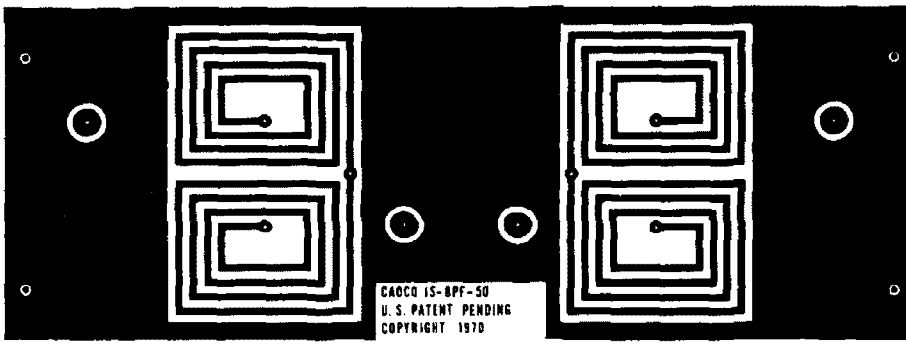
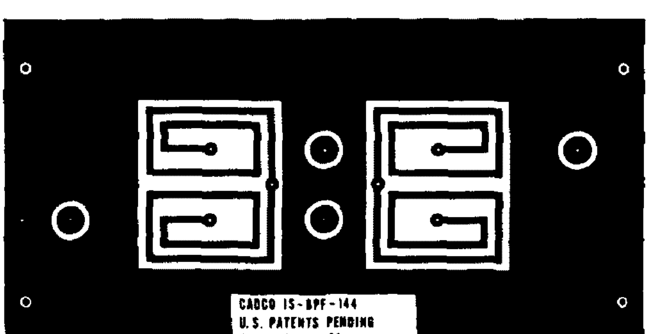


fig. 3. Half-size layout of the 50-MHz bandpass filter board.

The 144-MHz filter has a 3-dB bandwidth of 3.0 MHz, centered on 145 MHz. Insertion loss between 144.0 and 146.0 MHz is on the order of 1.0 to 1.5 dB, depending somewhat on the quality of the parts used in the filter. The bandpass skirts are very sharp, dropping down 3 dB at 143.5 and 146.5 MHz, and down 30 to 35 dB at 140 and 150 MHz.

In either of the filters the width of the passband may be designed to cover a much narrower range: 250 kHz at either 50 or 144 MHz. In this case, rejection 3

fig. 4. Half-size layout of the 144-MHz bandpass filter board.



parts selection

Capacitors C1, C3, C4 and C6 determine the width of the passband as well as playing a part in establishing the operating frequency of the filter. Ceramic discs are adequate for these capacitors, provided leads are short, with direct point-to-point wiring. The capacitance values are very important. If a 2.7 pF ceramic disc is specified, only 2.7 pF will work – 2.5 or 3.0 pF *will not.* Five percent capacitors recommended.

Inductors L1 and L3 are the fixed etched inductances. L2 is an airwound inductance used for fine adjustments. As L2 is varied from maximum to minimum inductance, the operating frequency moves upward from 120 to 165 MHz.

*Complete boards, as well as complete parts kits with boards, are available from CADCO, Suite 107, 4444 Classen Boulevard, Oklahoma City, Oklahoma 73118. 50-MHz bandpass-filter board, $6.00; 50-MHz bandpass-filter kit, BPF-50, $11.00; 50-MHz preamp board, $8.00; 50-MHz preamplifier kit, IPA-50, $19.50. 144-MHz bandpass-filter board, $6.00; 144-MHz bandpass-filter kit, BPF-144, $11.00; 144-MHz preamplifier board, $8.00; 144-MHz preamplifier kit, IPA-144, $19.50. All prices are postpaid in the U. S. A. If you order a complete kit and want the narrow-bandwidth version, simply specify "narrow band;" no price change.

Correct values of C1, C3, C4 and C6 vary from approximately 2.9 pF at 120 MHz to 2.4 pF at 165 MHz. If the values of C1, C3, C4 and C6 are too *large,* the width of the bandpass window will be too great for amateur applications; if the capacitance values are too small, the passband will be too narrow, and the insertion loss will be unnecessarily high.

Variable capacitors C2 and C5 are Arco-Elmenco trimmers; the printed-circuit board has been designed with their physical dimensions in mind.

The coaxial loop between C3 and C4 is a short length of RG-58/U coaxial cable. *Do not* try to replace this loop with another type of coaxial cable, or with a low-value coupling capacitor.*

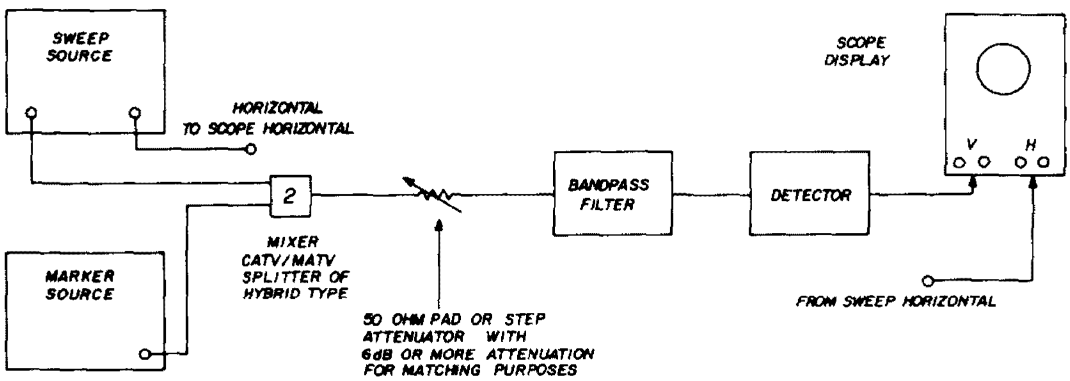


fig. 5. Test setup for aligning the etched-inductance bandpass filters.

tuning the bandpass filter

The etched-circuit bandpass filter may be aligned with a sweep generator and marker oscillator, or with a simple signal source (signal or marker generator).

Sweep generator/markers. Connect a broadband sweep generator to the input of the filter through a mixer device as shown in fig. 5. The mixer combines the sweep and marker signals into one composite signal that drives the filter (the mixer is not required for sweep generators which use an internal marker oscillator). The output of the filter is fed to a detector that demodulates the signal for the oscilloscope display.

Put the marker on the desired *center* frequency and tune C2 and C5 for maximum signal at the center of the passband. Now move the marker to the predicted 3-dB down point on the low side of the passband, and adjust C2 for maximum rejection. Your center frequency point *should not move up or down in frequency* as you do this.

Finally, move the marker to the predicted 3-dB down point on the high *side* of the passband, and adjust C5 for maximum rejection. Again, the center of the passband *should remain centered* on the display.

Disconnect the sweep generator; with the marker at the center frequency of the filter, feed the marker signal directly into the filter. Connect the output of the filter to your receiver and read the level of the marker signal on the S-meter. (It is advisable to attenuate the marker output so the S-meter indication is in the range of S5 to S7, normally the receiver's most linear region.) Now connect the marker oscillator to the receiver and read the S-meter. The difference between the two S-meter readings is the insertion loss of the filter. If everything is adjusted properly, the 1.0- to 1.0-dB insertion loss won't even be noticed on the S-meter.

Marker or signal generator. The alignment process with a marker or signal generator is essentially the same as that with a sweep generator. First, tune the filter for maximum signal at the center passband

*For 75-ohm systems the RG-58/U coaxial cable may be replaced with a section of RG-59/U. No other circuit changes are necessary.

frequency. Check filter and no-filter S-meter readings as you progress to see when you get down to the level of 1- to 2-dB insertion loss. Move the signal generator to the lower 3-dB down frequency and adjust C2 for maximum rejection (go back to the center frequency to make sure it hasn't moved too). Now move the generator up to the upper 3-dB down frequency, adjusting C5 for maximum rejection. Finally, check insertion loss to make sure it is still less than 2.0 dB.

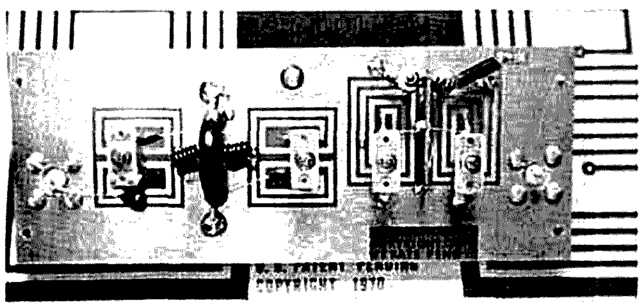
This bandpass filter arrangement uses fewer components, no shielded component sub-sections, and tunes up much easier than any other design that I am aware of. If it appears that you are having trouble holding the center passband frequency in place as you vary C2 and C5, try spreading a few turns of L5, and returning to the center frequency. (Spread the center of L5 apart about one extra turn width to start.) The change in L5 will compensate for slight unbalances between the twin sections of L1-L3 and L4-L6 that may exist because of differences between the fixed ceramic capacitors.

two-meter etched-inductance preamp

The two-meter etched-inductance preamplifier shown in the photograph consists of an etched-inductance bandpass filter and a single-stage grounded-gate fet amplifier. The circuit is shown in fig. 6. At 144 MHz the gain of this device is 10 to 12 dB, and the noise figure is 1.4 to 1.6 dB. A similar unit for 50 MHz will produce 10 to 12 dB gain with a noise figure of 1.0 dB of less.* Half-size circuit-board layouts for 50 and 144 MHz are shown in figs. 7 and 8 respectively.

A Siliconix 2N5397 fet was used in the original interdigital preamplifier described in the August issue.[2] Since that time Siliconix has introduced a low-cost plastic version of this device, the E-300. The price of the new E-300 is $2.00 in small quantities, as opposed to $5.50 for the 2N5397.

In the photograph you can see that the input inductance to the fet (L7 in fig. 5) is isolated from the output inductance (L8 in fig. 5) by a piece of double-sided copper-clad printed-circuit board. In addition to serving as a shield, this section of board is a mounting plate for the E-300 fet. The shield is soldered to the 1/8-inch strip of copper between L7 and L8; the E-300 gate lead is soldered to the L8 side of the shield with the transistor mounted inside a 3/8-inch mounting hole (see fig. 9).



144-MHz interdigital preamplifier. The filter section is to the left side of the board, the fet preamplifier to the right.

Capacitor C6 feeds the input of the preamplifier. On both versions, C6 is tapped onto L7 at the point indicated by the dot in fig. 7 and 8. C6 should be routed from the output end of L6 to the input of L7 under the printed-circuit board. This is the only part mounted under the board.

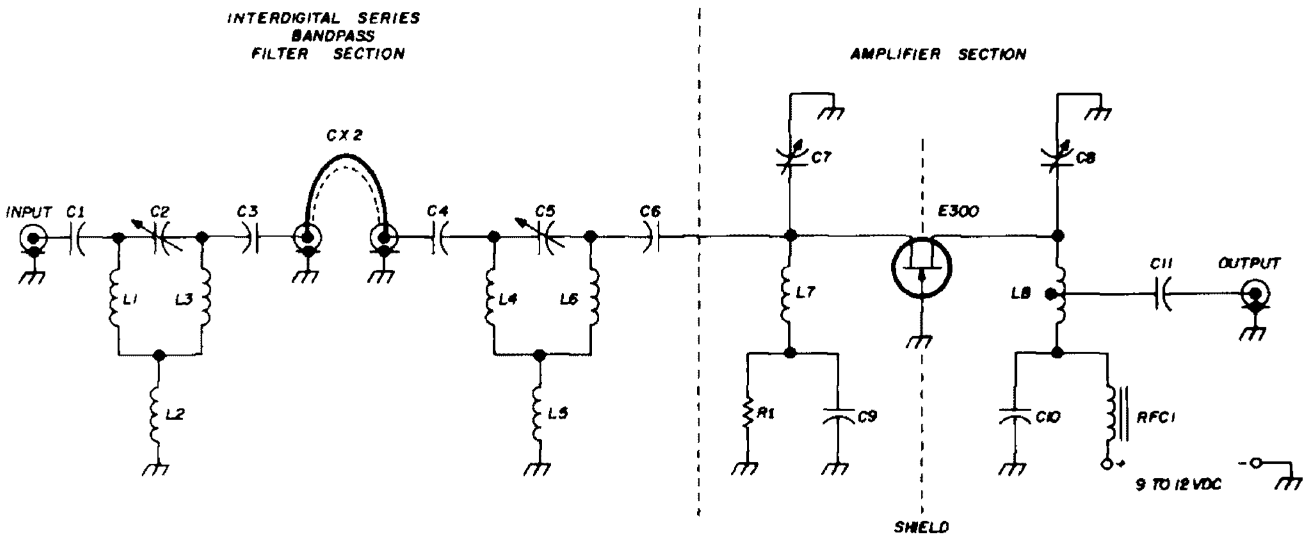
Capacitors C7 and C8 are Arco-Elmenco trimmers, and resonate with inductances L7 and L8. L7 and L8 are

*Since noise is wideband, a broad rf amplifier reacts to noise not only within the desired range of frequencies, but to noise outside that range as well. The steep skirted bandpass filter contributes measureably to an overall noise reduction in the communications system because the noise seen by the receiver is limited to that noise within the passband of the filter. As a consequence, the mixer is hit only with in-band noise, and while difficult to measure accurately, the receiver seems less prone to noise blocking.

bypassed to ground with BH-140 stud-type capacitors mounted at the outside end of the etched inductances. The BH-140 capacitors are mounted upside down, with the solder tab soldered to the end of the shield. Ceramic button bypass capacitors could also be used for this purpose.

Resistor R1 is chosen for correct fet operating current. This 1/4-watt resistor is mounted with very short leads from the outside end of L7 to ground, and placed against the shield at the C9 soldering point. The proper value for R1 is determined by placing a milliammeter in series with the dc supply (9 to 12 volts) and adjusting the resistance for 5-mA drain current; this is the proper current drain for both minimum noise figure *and* maximum gain. The correct value usually falls

fig. 6. Etched-inductance interdigital preamplifier for 50 or 144 MHz.



50-MHz bandpass preamplifier

C1, C3, C4, C6 6.0 pF, 5% disc or tubular ceramic capacitor

C2, C5, C7, C8 4 — 40 pF midget trimmer (Elmenco-Arco type 422)

C9, C10 500-pF stud-type uhf bypass capacitors (Sprague type BH-140)

C11 500-pF disc ceramic capacitor

L1, L3, L4, L6, L7, L8 etched inductances on printed-circuit board (fig. 8)

L2, L5 15 turns no. 16 solid copper, close-wound on 1/4-inch form, with last two turns (ground end) separated from balance of coil by one turn width

R1 91 to 560 ohms, 1/4-watt (see text)

RFC1 Ohmite Z-50 rf choke (7.0 μH) or equivalent

CX 2-1/2 inches RG-58/U coaxial cable, with braid twisted into pigtails so that total length of inner dielectric is 1-3/4 inches; pigtails are 1/2-inch long

144-MHz bandpass preamplifier

C1, C3, C4, C6 2.2 pF, 5% disc or tubular ceramic capacitors

C2, C5, C7, C8 4-40 pF midget trimmers (Elmenco-Arco type 422)

C9, C10 500-pF stud-type uhf bypass capacitors (Sprague type BH-140)

C11 500-pF disc ceramic capacitor

L1, L3, L4, L6, L7, L8 etched inductances on printed-circuit board (fig. 9)

L2, L5 5 turns no. 16 solid copper, close-wound on 1/4-inch form with 0.2 inch standoff pigtails

R1 91 to 560 ohms, 1/4-watt (see text)

RFC1 Ohmite Z-144 rf choke (1.8 μH) or equivalent

CX 2-1/2 inches RG-58/U coaxial cable, with braid twisted into pigtails so that total length of inner dielectric is 1-3/4 inches; pigtails are 1/2-inch long

between 100 and 560 ohms, with 200 to 400 ohms being common.

RFC1 is an Ohmite Z-144 or Z-50; although their current-carrying capacity isn't required for this application. Any good quality wirewound rf choke with the same inductance value as the Z-50 (7.0 μH) or Z-144 (1.8 μH) will do. Since L8 is bypassed to ground with the 500-pF BH-140 there's not much chance of an rf problem anyway, but it's good practice to include the rf choke in the circuit.

The output coupling capacitor is a 500-pF disc ceramic. This capacitor is tapped onto L8 fairly close to the point where C10 is attached to the outer end of the inductance. A small dot on figs. 7 and 8 indicates the approximate tap point for maximum preamplifier gain.

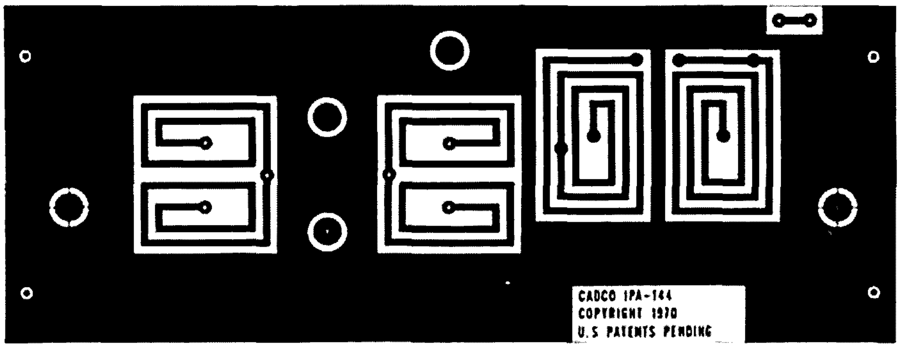
preamplifier tuneup

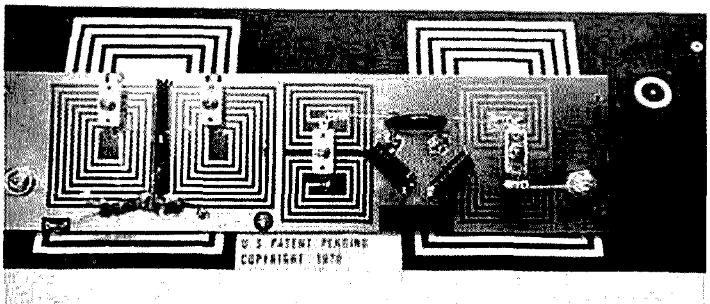
When aligning the preamplifier, the bandpass filter section must be aligned first. To accomplish this, the transistor stage must be temporarily eliminated from the circuit. An extra coax connector mounting hole (see figs. 7 and 8) is provided for this purpose. Install a coaxial fitting as indicated, route C6 to the fitting and align the filter as previously described. When the bandpass filter is properly aligned, transfer C6 from the extra coaxial fitting to L7 and remove the fitting from the board.

Apply voltage (9 to 12 Vdc) to the preamplifier and select R1 for 5-mA drain current. Apply an input signal — at the center frequency of the bandpass filter — and tune C8 for maximum indicated level on the receiver S-meter. Keep signal level down so the meter reads in the range from S5 to S7. With C8 peaked for maximum signal, tune C7 for maximum. C7 will tune somewhat more broadly than C8. Check the setting of C8 again, keeping the output of the rf signal source at a relatively low level.

Disconnect the filter-preamplifier and measure the output of the signal generator with your S-meter. (It is assumed that you have a 50-ohm-output signal source, or can rig one with your antenna system and a grid-dip oscillator across the yard.) Put the filter-preamplifier back in the system and measure the output level with

fig. 7. Half-size layout for the 144-MHz bandpass-filter preamplifier board.





50-MHz interdigital preamplifier. The filter section is to the right; fet preamplifier to the left. Capacitor C6 is mounted under the board.

your S-meter. There should be 10 to 15 dB additional signal level with the preamplifier; noise level, with the antenna disconnected, should be noticeably less than your present converter.

The total gain of the preamplifier is the gain of the E-300 grounded-gate stage, less bandpass-filter losses. Therefore, although the E-300 is capable of 15 dB or more gain at 50 or 144 MHz, filter losses of 1 to 2 dB reduce total package gain to 10 to 12 dB.

The grounded-gate preamplifier is unconditionally stable. The isolation between *wells* (term applied to the etched inductances on the printed-circuit board) has been measured as high as 70 dB with as little as 1/8 inch copper between wells. In commercial CATV preamplifiers, for example, I have run four cascaded fet stages with filters on a single board with no instability; with four stages voltage gain is on the order of 40 dB.

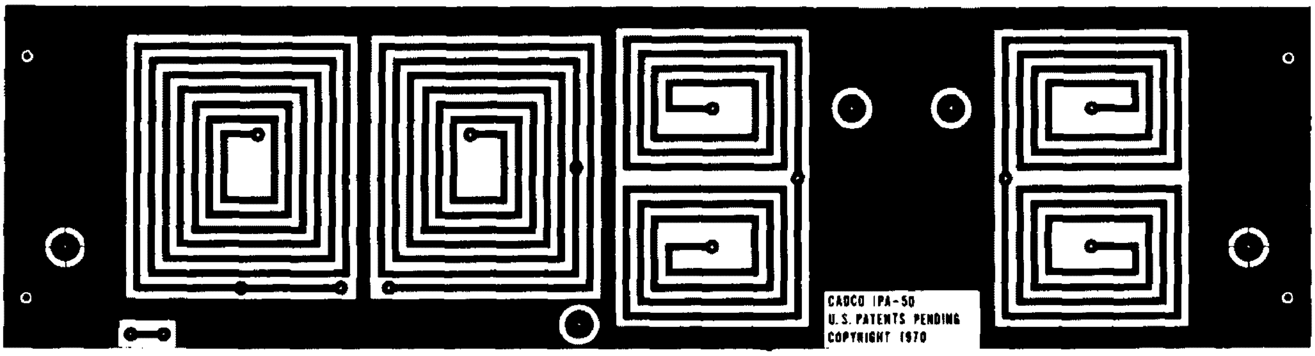
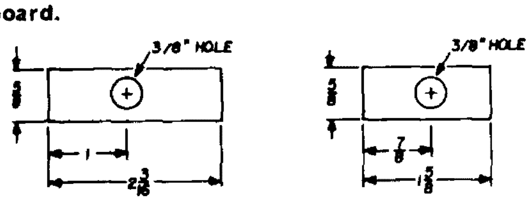


fig. 8. Half-size layout for the 50-MHz bandpass-filter preamplifier board.

fig. 9. Layout for the E-300 mounting shields. Cut out from double-sided G10 printed-circuit board.



summary

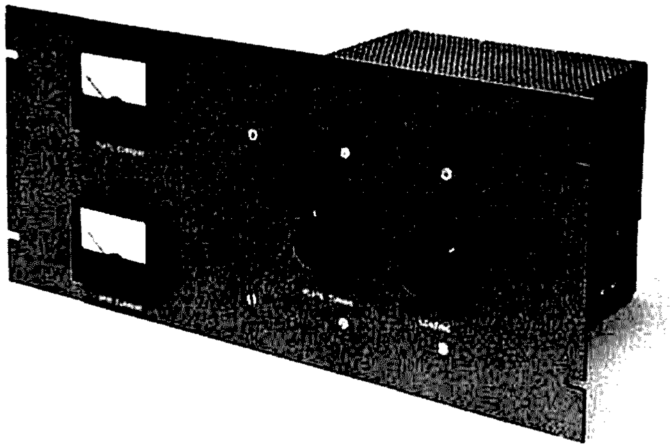
The subject of etched inductances for vhf receiver systems has been barely scratched in this article, as I am painfully aware. Additional prototype work has been done in other areas including an etched-inductance converter (with bandpass filter), etched-inductance transmitting mixers, etched-inductance hybrid couplers, and etched-inductance stopband filters (with preamplifier) to replace repeater cavities.

Etched-inductance circuits are relatively simple projects for the serious experimenter. Simply use 1/8-or 1/16-inch tape* to form the inductances, with 1/16 inch between inductances. Etched inductances are measured by the square well area, and a 1.25-square-inch well with 1/16-inch inductors and 1/16-inch spacing between inductors hits 144 MHz. The etched inductances can be easily tapped, and you can re-tap many times if you use a 25- or 35-watt iron and don't use sustained heat on the etched strips.

references

1. Charles Ryder, U.S. Patent number 1,837,678, "Inductance Coil — Particularly Adapted for use with Radio Tuning Devices," December 22, 1929.
2. Robert Cooper, Jr., W5KHT, "Interdigital Preamplifier and Comb-line Bandpass Filter for VHF and UHF," *ham radio,* August, 1970, page 6.

ham radio



two-kilowatt linear amplifier for six meters

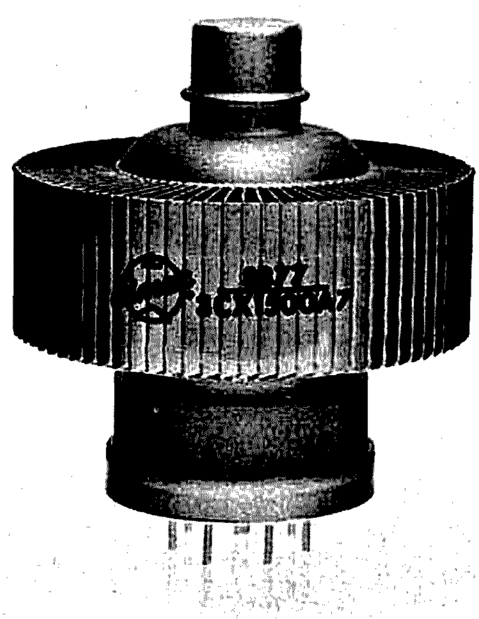
This high performance six-meter linear features the new Eimac 8877 and provides excellent stability, good reliability and minimum harmonic output

Robert I. Sutherland, W6UOV, EIMAC Division of Varian, San Carlos, California 94070

The serious six-meter operator needs a high power amplifier that will function reliably over extended periods of time and have minimum harmonic radiation. Such amplifiers seem to be commonplace for the "dc bands" but are rather rare for 50 MHz and above. Many six-meter amplifier designs are cranky, hard to neutralize or otherwise unstable or tricky to adjust.

The amplifier described in this article has none of these undesirable attributes. It will run key-down on a 24-hour basis, if need be, and is stable and easy to adjust. I have used it over a period of months and it has proven to be a valuable

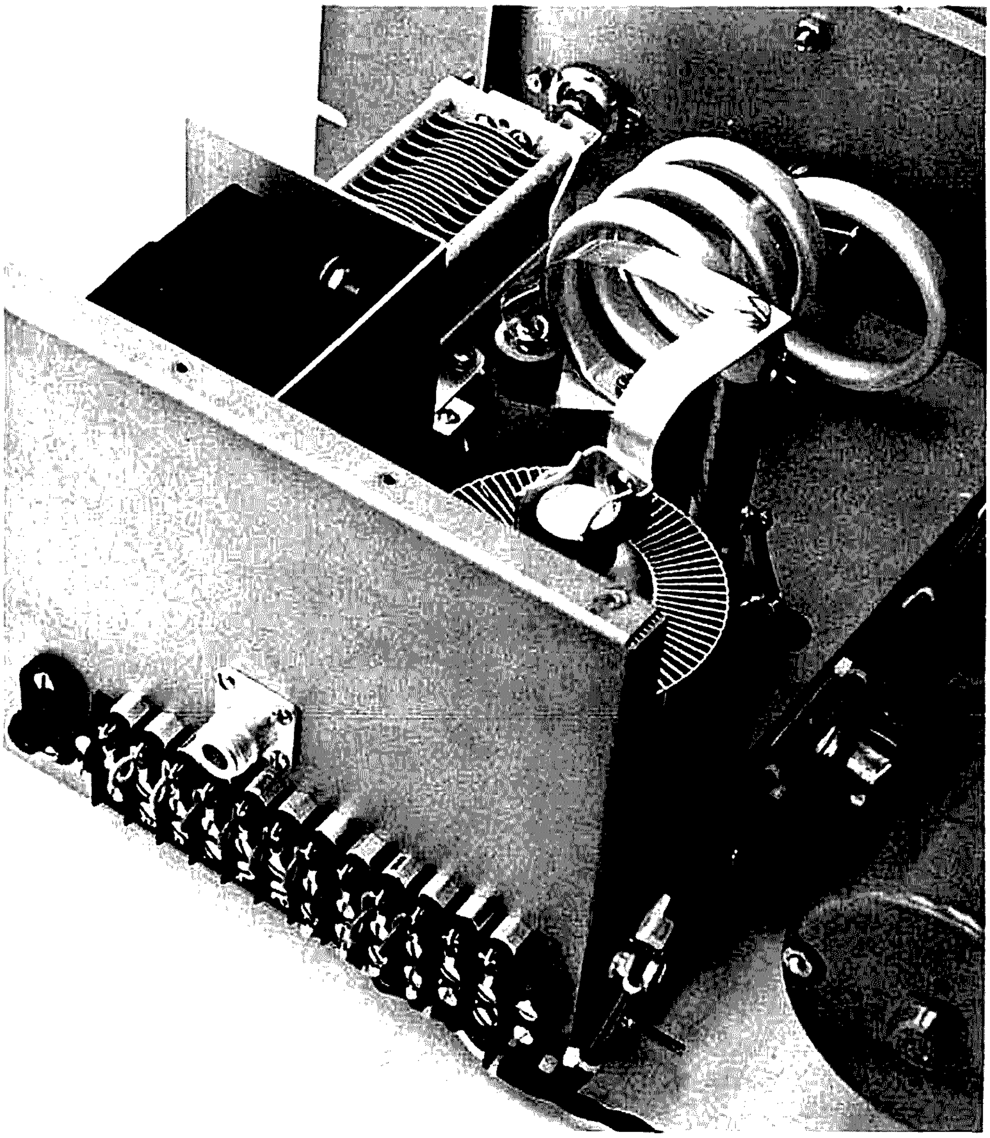
The new high-mu 8877/3CX1500A7 triode recently announced by EIMAC.



adjunct to the spread of six-meter equipment in my station.

This amplifier uses a grounded-grid circuit with a new high-mu triode just announced by Eimac: the 8877/3CX1500A7. This ceramic/metal triode is intended for linear service in the high-frequency and vhf range. The amplifier is intended for the maximum legal power input, 1000 watts dc, and can develop up to 2000 watts peak envelope power input during ssb operation. The amplifier requires a driver that can supply approximately 40 watts PEP at 50 MHz. Using a plate potential of 2600 volts and plate current of 750 milliamperes, power output will be about 1200 watts. This represents an amplifier efficiency of 61% and a power gain of 14.8 dB.

A schematic of the amplifier is shown in fig. 1. The control grid is operated at dc ground with a minimum of inductance between the tube and the chassis. The plate and grid currents are measured in the cathode return lead. A 12-volt 50-watt zener diode is placed in series with the cathode return lead to set the desired idling plate current. No special neutralization scheme is needed to attain completely stable operation.



Top view of the plate circuit of the linear amplifier showing the shorted-turn tuning scheme. The shorted-turn is hard-soldered to shaft coupler to allow front panel tuning. The "anti-inductance" strap can be seen connecting the top of the plate choke to the plate blocking capacitor. Note that the position of the plate blocking capacitor can be changed by loosening one screw and rotating the capacitor around the screw.

The plate circuit is a standard pi-network with tube output capacitance plus stray capacitance to the cabinet forming the input capacitance of the network (30 pF). The output loading capacitor is an air variable shunted by two fixed ceramic power. The input impedance of the tube is 54 ohms resistance in parallel with 26 pF capacitance. The match holds over the 1-MHz tuning range of the amplifier.

A 10,000-ohm 25-watt resistor in the cathode lead of the 8877/3CX1500A7 is

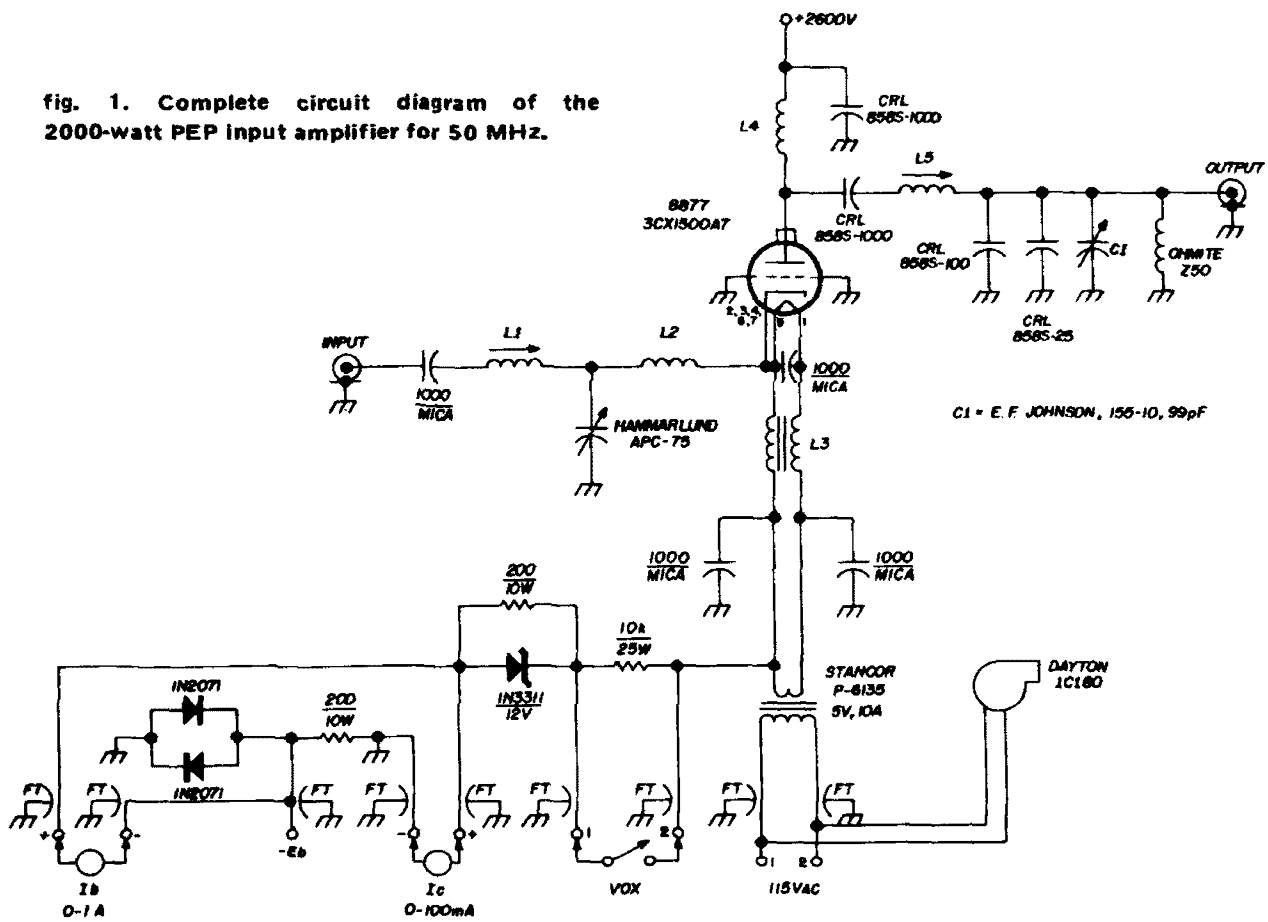


fig. 1. Complete circuit diagram of the 2000-watt PEP input amplifier for 50 MHz.

L1 6 turns no. 18 on a CTC 1538-4-3 form; coil length 7/8"

L2 6 turns no. 18, ½" diameter, 5/8" long, self-supporting

L3 Bifilar wound choke, ½" diameter core, 3" long, each coil 12 turns no. 10 Formvar; core is Indiana General CF-503

L4 54 turns no. 20 enameled on 1/2" diameter Teflon rod; winding length 1-13/16"

L5 3 turns 3/8" diameter copper tubing; inside diameter 1-7/8"; coil length 2-3/8"; shorted turn 2-¼" diameter 3/8" copper tubing ¼" from main coil

FT Erie 327 1000-pF feedthrough capacitors

capacitors. Amplifier tuning is accomplished by varying the inductance of the coil by adjusting the coupling between the coil and a shorted turn.

The cathode input circuit consists of a simple T-network. The network was calculated so that a 50-ohm cable from the driver would be matched to the input impedance of the 3CX1500A7 at full used to reduce standby current through the tube to a low value. When the exciter is turned on, a set of contacts on the vox relay (or other control relay) shorts out the 10,000-ohm resistor, causing the tube to operate at its normal idling plate current. The 200-ohm 10-watt resistor from the negative terminal of the plate supply to ground makes certain the nega-

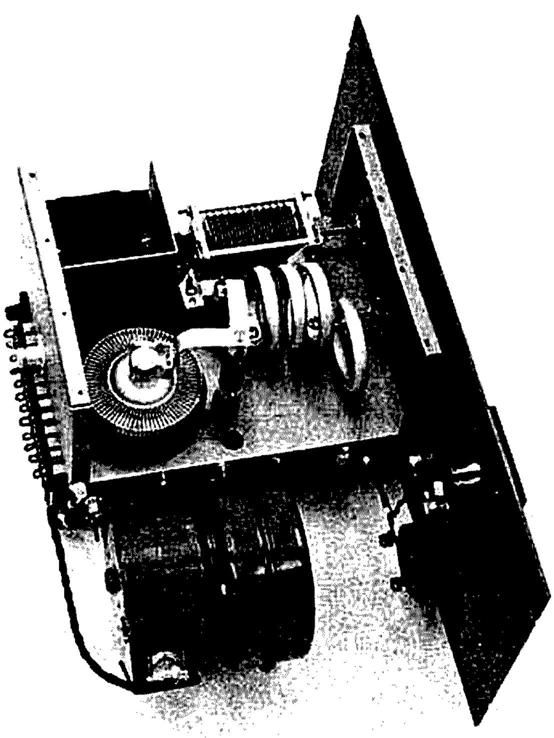
tive terminal does not soar to the value of the plate voltage if the positive side of the power supply is accidentally shorted to ground.

The two 1N2071 diodes across the 200-ohm resistor limit any transient surges under the shorted condition which might cause insulation breakdown. Also, these diodes afford some transient protection of the two meters by providing a path around the meters. Additional protection could be obtained by putting two back-to-front parallel connected diodes across each meter. The 200-ohm resistor around the zener provides a load for the zener and prevents the cathode voltage from becoming quite high if the zener should burn open.

the plate circuit

Top views of the amplifier chassis are shown in the photographs. The closed ring near the front panel is the shorted turn used for tuning; it is made of 3/8-inch diameter tubing, hard soldered to a brass shaft coupler with copper-silver solder. Soft solder would not be advisable in this application because of the high circulating current in the shorted turn. The "anti-inductance" strap is used to set the tank circuit to the desired tuning range. This strap runs from the top of the plate rf choke to the plate blocking capacitor. The position of the blocking capacitor can be moved to allow the strap to be flexed and set to the proper position. Note that the current through the strap is going in the opposite direction from the current in the coil at any instant and therefore causes field cancellation.

Another view of the plate circuit. The air variable across the top edge of the chassis is the adjustable part of the loading capacitor. Two ceramic barrel capacitors are mounted in parallel with the air capacitor and can be seen at the end of the variable capacitor near the filament transformer shield.



To set the amplifier to the low-frequency end of the band, the shorted turn is completely decoupled and the position of the blocking capacitor and the anode strap adjusted to resonate the plate circuit to 50 MHz. As the shorted turn is coupled tighter, the total inductance in the plate tank circuit will be reduced, causing the resonant frequency to increase. When the shorted turn is fully coupled, the resonant frequency of the plate tank circuit will be about 51 MHz.

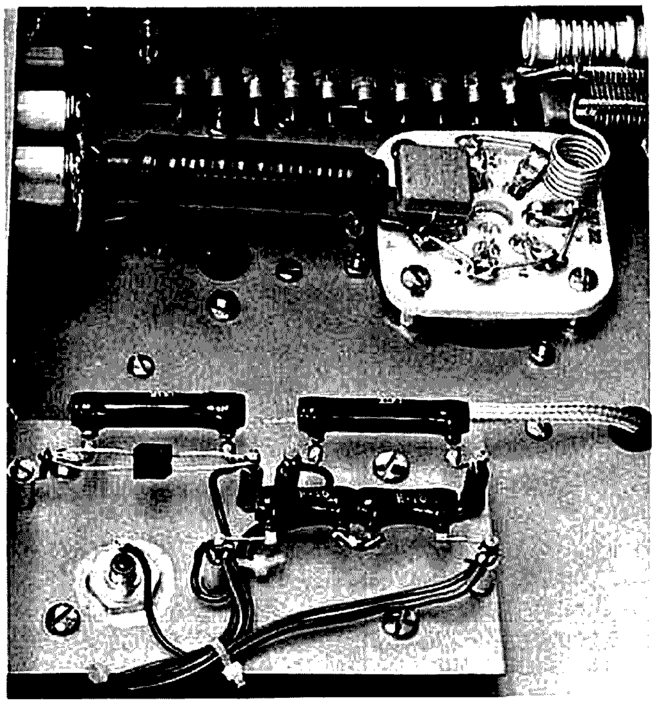
Amplifier loading is accomplished in the same manner as in a typical pi-network amplifier. The loading capacitor is the air variable along the top right edge of the chassis. The two ceramic fixed capacitors are at the left end of the air capacitor and at the end of the coaxial cable coming from the type-N coaxial receptacle mounted on the back panel.

The plate choke is made of 54 turns of no. 20 enameled wire closewound on a one-half inch diameter Teflon rod. The winding length of the coil is 1-13/16 inches. The choke is mounted on top of the ceramic capacitor which is used to by-pass the B-plus end of the choke.

Visible on the back of the front panel are the *Jackson* ball-drive assemblies. These handy devices provide a very smooth and slow "feel" to the tuning. The 5.0-volt 12-ampere filament transformer is visible inside its aluminum shield at the top left end of the chassis.

the input circuit

The input matching network is a standard T-design consisting of two series coils and one shunt capacitor. One coil and the shunt capacitor are variable. With these two adjustments it is possible to



View of the underside of the chassis showing the input circuit and the location of the zener diode and resistors. The T matching network is in the upper right hand side of the chassis. The heater-cathode choke is mounted between the socket and the ceramic stand-offs at the left side of the picture. Note that the socket is mounted below the chassis to allow passage of the cooling air. The straps grounding the grid to the chassis can also be seen under the threaded brass spacers used to sub-mount the socket.

cover a wide range of impedance transformations. The controls for the variable elements are brought out the left rear side of the chassis. Once the adjustments have been made, no tuning is required over the first megahertz of the band.

The input matching network can be seen in the top right corner of the under chassis photograph. The cathode-heater rf choke is near the tube socket. The choke is bifilar wound with twelve turns on each

*Available from Newark Electronics Corporation, 500 North Pulaski Road, Chicago, Illinois 60624. Order catalog number 59F1521, $2.50 plus shipping.

coil using no. 10 Formvar insulated wire. The core material is *Indiana General CF-503,* one-half inch in diameter.* The core permeability is a little high for this application, but the material was available and has not given any trouble. The Johnson 122-247-202 socket is mounted one-half inch below the chassis using threaded brass spacers. Four pieces of brass shim stock, or beryllium copper, are formed into an "L" shape to mount between the brass spacers and the chassis and make contact to the control grid ring.

the tube

The 8877/3CX1500A7 is a new ceramic triode having good division between the plate current and the grid current. It has EIA base no. E7-2 which can be used with the standard septar sockets. The tube has a plate dissipation rating of 1500 watts, and has a mu of approximately 200. The cathode is indirectly heated, and the filament requirements are 5.0 volts at 10 amperes.

performance data

Many different operating conditions were tried with this amplifier. The conditions most suitable for amateur ssb operation at 2000 watts PEP input are:

Plate voltage	2600 Vdc
Plate current (single-tone)	750 mA
Plate current (idling)	40 mA
Grid voltages	-12 Vdc
Grid current (single-tone)	58 mA
Power input	1950 W
Power output	1200 W
Efficiency (apparent)	61 %
Drive power	40 W
Power gain	14.8 dB

The intermodulation distortion products at full peak envelope power input under the above operating conditions are:

3rd order	-44 dB
5th order	-37 dB
7th order	-64 dB
9th order	-68 dB

ham radio

speech clipping

in single-sideband equipment

Audio speech clipping produces excessive distortion when used with ssb equipment — here's why

Walter Schreuer, K1YZW, Riverbank, Ipswich, Massachusetts 01938

In the old days when ssb was unknown on the amateur bands the intelligent use of a speech clipper with a carefully designed filter frequently made contacts possible which otherwise could not have taken place. Distortion was quite bad, so much so that the readability "in the clear" was actually degraded slightly, but you could – and did – switch the device out of circuit when conditions were good. Unfortunately, when you try to use the same simple technique with ssb the results are disappointing, often atrocious. Many amateurs have even abandoned the scheme.

Too much confusion still exists regarding speech processors. For example, a completely illogical comment in a well-known British magazine recommended speech clipping to avoid overdriving the transmitter output stage while recognizing that the process is deficient in other respects. I recall the outburst of an exasperated ham who received consistent adverse reports on his $15 transistor clipper: "It must work – you can't be listening right!"

Actually, there are several references that point out the incompatibility of speech clipping with the ssb. An early edition of the SSB Handbook (put out by Collins Radio) is quite outspoken on the subject. The following is an attempt to shed some light on the matter in simple terms, at the risk of over-simplification, and to briefly examine some preferable alternatives.

some basic facts

The human voice has a low average-to-peak power ratio; our transmitting apparatus is limited primarily to a specific peak power level. If this power is exceeded, the result is excessive distortion and bandwidth, among other possible effects. In a-m and fm systems the situation can be improved, at the expense of fidelity, by instantaneously peak limiting or clipping the audio frequency signal prior to the modulation process. Considerable harmonic and intermodulation distortion products are generated by the clipper which can greatly increase the

bandwidth of the signal. For this reason the clipper is always followed by a low-pass filter designed to cut off at 3kHz or so. To a small extent this filter negates the action of the clipper, but even so, effective gains of 10 dB or higher can be obtained. (Please note this applies to a-m and fm, *not* ssb.)

It may seem strange to talk about gain in connection with devices intended to

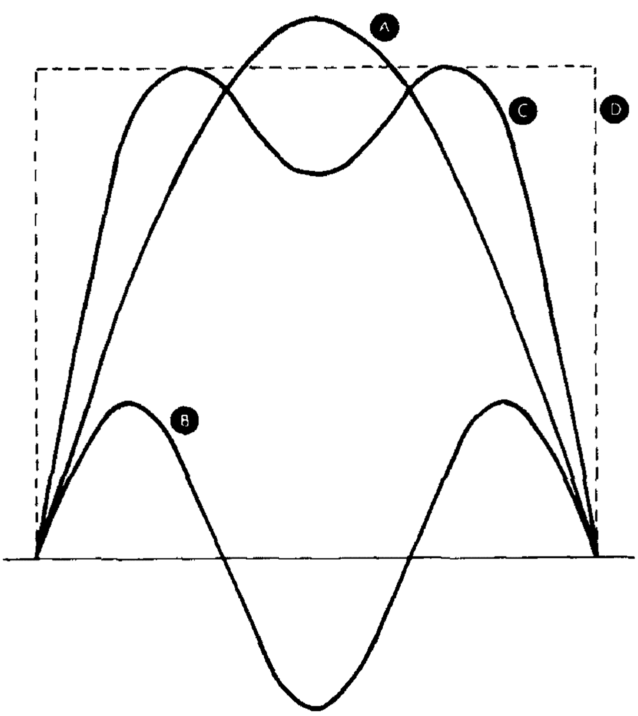


fig. 1. The derivation of a square wave. The fundamental component is shown in A. B is the third harmonic. The sum of A and B is shown in C. D is the resultant with higher odd harmonics of proper amplitude and phase.

improve readability without increasing the peak power output. However, the definition is quite simple: the gain is the ratio of the peak power of an unprocessed transmission to that of a processed one, where the former produces the same readability as the latter at the distant receiver under very poor conditions.

We also talk about the amount of clipping in decibels. This is simply a measure of the reduction of the largest speech peaks, or more precisely, the ratio of the gains of the audio amplifiers with and without clipping for the same peak power output. In a-m or fm links 15 dB clipping can result in gains of 9 to 12 dB. This produces the same readability at the distant end, under poor conditions, as an unclipped transmission with 8 to 16 times peak power.

phase relations

When you severely clip a sine wave or single-tone signal, the resultant approximates a square wave. This can be thought of as the combination of another sine wave and its odd harmonics. The latter have definite relationship to the fundamental frequency component in both amplitude and phase. **Fig. 1** shows a sine wave **(fig. 1A)**, a third harmonic component **(fig. 1B)** and their resultant **(fig. 1C)**. If you add more odd multiples in proper amplitude and phase you eventually end up with a square wave **(fig. 1D)**.

If the phase of the harmonic signals is shifted so that their peak amplitudes add to that of fundamental tone the results would be as in **fig. 2**. The peak amplitude of the resultant **(fig. *2D*)** is more than twice (8 dB greater) the square wave of **fig. 1**. If the peak amplitudes of all the harmonics subtract from the fundamental the dip in the middle of the resultant waveform would approach zero! These examples of extreme cases of phase shift emphasize their importance.

ssb generation

In an ssb exciter the phases of the individual components of the audio signal are not maintained. The balanced modulator does no harm in this regard. The *resultant* amplitude of the upper and lower sidebands forming the dsb signal remains peak limited in a properly designed and adjusted circuit with clipped audio input. (One point in favor of dsb is that speech slipping *can* be used.) However, as soon as one sideband is removed the delicate balance of the component signals is upset, resulting in severe amplitude variations. You are left with a set of rf components which correspond in amplitude and frequency to the audio fundamental, its harmonics generated by the clipping action. At rf these com-

ponent signals are not harmonically related, and with the carrier and opposite sideband removed add up arithmetically to form the composite signal. Periodically the rf envelope exhibits maxima and minima with magnitudes dependent upon modulating frequency and system bandwidth.

As a practical example, assume that we have a heavily clipped (approximately square) 400-Hz wave of 0.8 volt amplitude and use it to produce a 100-kHz ssb signal. Also assume that conversion gain is unity. Table 1 shows the frequency components of the ssb signal and their amplitudes. Components corresponding to harmonics above the seventh are ignored since they are attenuated by the filter following the clipper. Periodically, all the peaks will coincide, resulting in a maximum of the rf envelope that is equal to the sum of the individual peak values. There will also be times when the peaks of each of the distortion components will coincide to subtract from the amplitude of the main component and give a minimum value of the envelope. In table 1 it is shown that the amplitude of the resultant envelope will vary 14 dB at an audio rate,* though a constant input signal is used! This is exactly the reverse of the desired effect. Bear in mind that this extraneous amplitude modulation is quite independent of any variation of the input signal. While the 14 dB variation (rather less at higher audio frequencies) is probably less than the normal variation in a speaker's voice the accompanying distortion often offsets any small numerical gain in average power.

phasing-type exciters

The situation, as described so far, is directly applicable to phasing exciters. If you severely limit the low-frequency response below 1000 Hz in the speech amplifier so the low notes are clipped only lightly, some gain in average power output can be obtained, together with noticeable distortion and obvious lack of bass. I do not believe that any gain realized through speech clipping in this manner is any greater than that obtainable with a properly designed alc or volume compressor system. The latter of course produce negligible distortion and can be designed to give properly balanced frequency response.

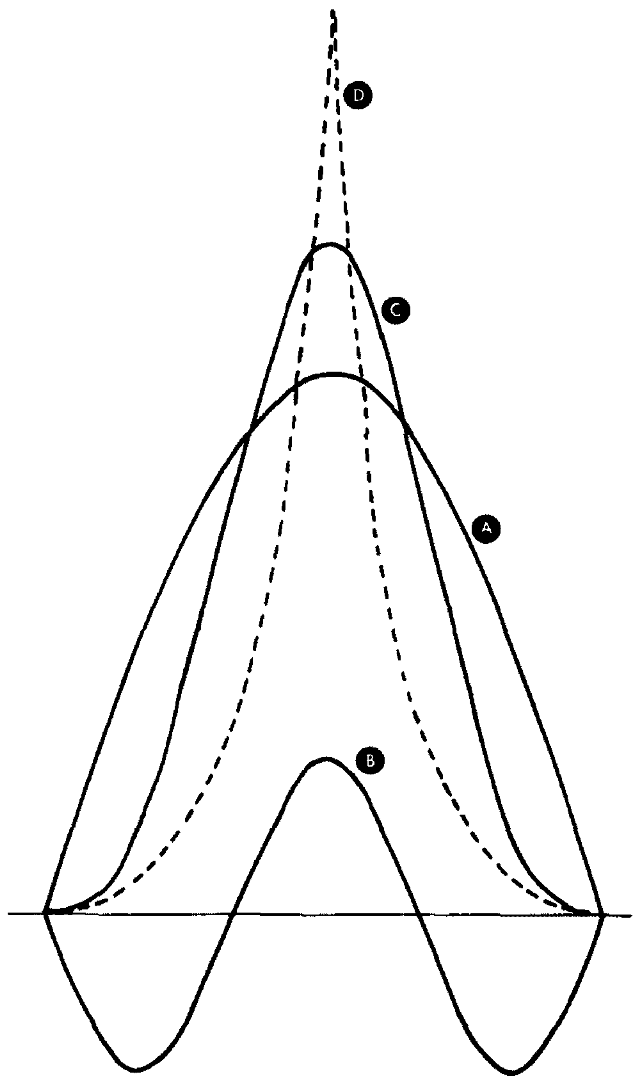


fig. 2. The affect of improper phasing of components. A is the fundamental; B is the third harmonic; C is the sum of A and B; D is the resultant.

filter-type exciters

The effects of speech clipping in ssb exciters are small compared to the further deterioration caused by the filters used in most modern equipment. The difficulty is the "transient" response of these filters,

*It can be shown that the amplitude modulation frequency is 800 Hz, or more generally, twice that of the input speech signal.

or the response to impulses or very rapidly changing rf envelopes. The effect is the same as that observed in receivers when impulse interference is present. Under those conditions you are often better off to switch out the ssb filter if no adjacent-channel interference is experienced.

A rapid change in the rf envelope causes the filter to give out a high amplitude spike followed by several cycles of damped oscillations. The initial spike is probably inaudible in your receiver because of its short duration. However, subsequent cycles constitute what is known as filter ringing. The amplitude of the spike, the ringing frequency and amplitude are a function of the filter; The effect worsens as the filter cut-off characteristic is improved.

table 1. Results when a severely clipped 400-Hz wave, amplitude 0.8, is used to produce a 100-kHz ssb signal.

	relative amplitude	af parameters	rf parameters
fundamental	1.00	400 Hz	100.4 kHz
3rd harmonic	0.33	1200 Hz	101.2 kHz
5th harmonic	0.20	2000 Hz	102.0 kHz
7th harmonic	0.14	2800 Hz	102.8 kHz
sum of distortion terms		na*	0.67
maximum amplitude		0.8	1.67
minimum amplitude		0.8	0.33
variation		0 dB	14 dB

*not applicable because of harmonic relationship

In ssb exciters appreciable speech clipping produces the same effect though some relief is afforded by the audio filter following the clipper. The ringing frequency is independent of the audio input to the exciter and represents additional and particularly vicious distortion. The initial spike can have a larger relative amplitude than the audio input which causes it, thereby requiring a reduction in audio gain below that which would have been used without the clipping device. In most cases the operator is unaware of this since his alc takes care of the problem; he finds it hard to believe that his audio is not only badly distorted but is often weaker than without the clipper!

I recall one case several years ago where the builder of a clipper reported that the only way he could make his gadget work properly was by pulling the alc rectifier tube in his Collins S-line. I am sure his ham neighbors must have objected in no uncertain terms; his results would have been less obnoxious if he had just removed the tube and omitted the clipper!

A practical demonstration of the nasty things I have described is not at all difficult. All it takes is an oscilloscope for viewing the envelope of the output of the transmitter and an audio square wave generator for the microphone input and for the sync input of the scope. The effects are best seen with a low frequency input between 300 and 500 Hz. Also instructive are the results obtained with a very low frequency input of 100 to 200 Hz, when only the harmonics can pass through the ssb filter. The output will resemble that of a badly distorted and overmodulated a-m wave. This demonstrates the absolute necessity for severe low-frequency filtering if a speech clipper must be used.

To conclude, speech clipping in any form is *incompatible* with the ssb mode as we know it. The generation of appreciable harmonics of the audio signal is the source of the trouble. Therefore, we must include as undesirable the many variations which produce the same effect: logarithmic limiters, instantaneous compressors, including those built into phone patches. I cannot recall all the names that have been used to describe essentially similar devices. They will do a good job on a-m or fm but not with ssb. Now let's stop being completely negative and take a look at schemes which can do a good or fair job of raising your talk power.

automatic level control

Basically, alc is intended to serve the same purpose as automatic gain control in receivers. One of the expressions must be a misnomer; the purpose in both cases is to obtain constant peak signal output by

controlling the gain of a preceding amplifying stage. A typical system is shown in **fig. 3.** When the peak rf signal exceeds the delay voltage the excess voltage appears across resistor R1. The rf is removed by a filter (R2 and C2) and the resulting dc voltage is used to reduce the gain of a preceding stage (preferably a stage operating at a different frequency in the interest of stability). The gain between the controlled stage and the sampling point should be large so that the actual peak rf level will exceed the delay voltage only slightly.

The attack time is largely determined by the product of R2 and C2, the charge constant. The charge time constant should be short so the duration of the initial rf peaks is brief and the resulting overmodulation (flat-topping) is of little consequence. Those familiar with servo loops will realize that the attack time

the output peak level nearly constant and equal to the delay voltage. Under this condition, which is typical of receiver agc circuits, no gain in average power is obtained. However, the benefits are considerable since the operator doesn't have to worry about overdriving the output stage. This can result in appreciable *psychological* gain.

If the alc recovery time is made very short – less than 50 mS or so – it becomes comparable to the periods of the lower audio frequencies and will result in distortion similar to flat-topping. A medium time constant – between 1/8 and 1/2 seconds – will avoid this and at the same time be low enough to compensate for the slow or syllabic amplitude variations of the human voice. This makes your voice sound a little unnatural, but no distortion nor increase in bandwidth will occur. The average power gain ob-

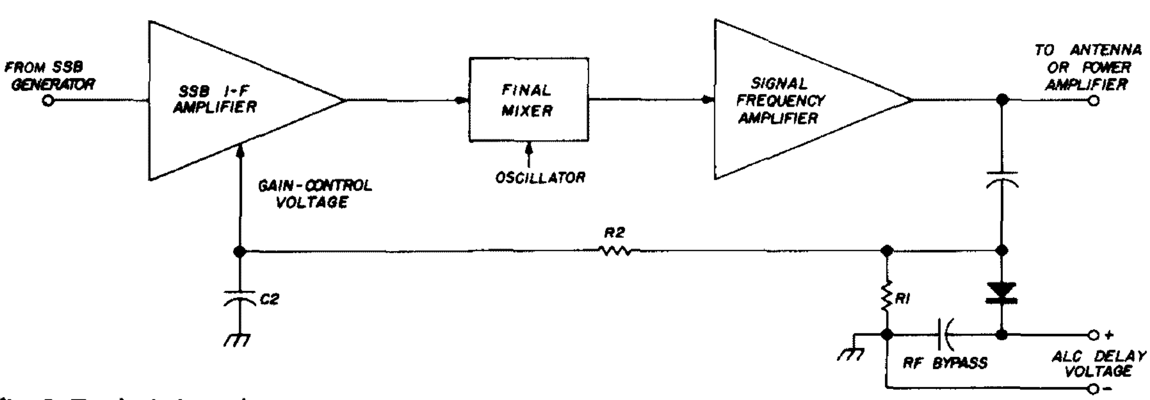


fig. 3. Typical alc system.

cannot be made too short or control overshoot followed by damped low frequency oscillations will take place. In practice there is no difficulty in obtaining a satisfactory compromise.

The decay or recovery time of the system depends on the product of C2 and (R1 + R2), the discharge time constant. Note that R1 does not affect the attack time; therefore, it may be selected to determine the recovery time only. If recovery time is long – several seconds – the output amplitude faithfully follows variations in the input signal with

tained by smoothing out the syllabic variations is between 3 and 6 dB in a well designed system and depends upon the operator's voice characteristics.

Alc is now a standard feature in modern ssb transmitters. Unfortunately, in many equipment its performance leaves a lot to be desired. In these cases alc is best regarded as an emergency brake and its extensive use is to be avoided. On the other hand, I know of one transmitter – and I am sure there are others – with excellent alc performance. All the operator has to do is turn up the

audio i-f level to obtain an essentially distortionless lift in average power.

It is not my purpose to enumerate the many dubious practices and glaring design errors I have noticed in some commercial amateur equipment. I once had to redesign and rebuild the alc circuitry in an exciter of well known manufacture before I was able to operate on the same band as its owner two miles away. His equipment was brand new and was functioning as designed but the signal was little better than it would have been when overmodulated without alc.

volume compression

Volume compression can be regarded as an agc system operating at audio. The performance is the same as a properly functioning alc system. In this case, too short a discharge time will result in speech clipping with its deleterious effects. A medium time constant (about ¼ second) will give an appreciable lift in average power.

If a volume compressor is used with a good alc system it will result in duplication with virtually no additional advantage. On the other hand, a good volume compressor is a considerable asset when used with exciters containing poor alc circuitry, or none at all.

rf clipping

Although rf clipping has come into prominence only recently it is an old idea. I have references which go back to 1952, and there is a patent dated 1926 for a similar process! Rf or i-f clipping will do for ssb what speech clipping does for a-m and fm, as well as quite a bit more. However, cost is rather high, and retrofitting older equipment is likely to be difficult.

As I have attempted to explain, the incompatibility of speech clipping and ssb stems from the audio frequency harmonics which are generated within the speech bandwidth. If the clipping process is postponed until after generation of the ssb signal the harmonics will be multiples of the intermediate or radio frequency and are nowhere near the fundamental. Their removal is no problem whatsoever, and you have a clipping process which is free from harmonic distortion. This is appreciably better than results with ordinary speech clipping in a-m or fm.

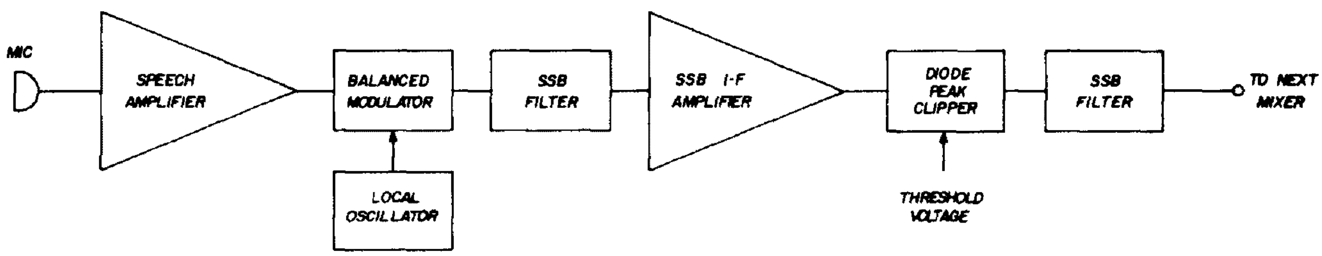


fig. 4. Single-sideband generation with rf clipping.

Unfortunately, there are intermodulation (IM) products: beat frequencies (sum or difference) when two or more signals exist simultaneously within the passband. Higher order IM products are the beat frequencies of the *harmonics* of the signals and these can fall in or near the fundamental signal band. The amplitudes of these IM products are fairly small as is their effect on signal quality. However, they may increase the bandwidth of the signal beyond acceptable limits. Therefore, a ssb filter is required after the rf or i-f clipper.

Fig. 4 shows a clipping arrangement operating at the first ssb frequency. The ssb generator is conventional, except that in a new design the first filter need only have moderate performance, say 20 dB sideband rejection and less carrier suppression assuming a good balanced modulator. The second ssb filter will determine

the quality of the outgoing signal and should have sharp skirts and a high attenuation floor. The agreement in the cut-off frequencies of the two filters must be very close on the carrier side (both sides if the same filters are used in a selectable sideband exciter). No overloading must occur in the stages preceding the i-f clipper. These requirements make modification of an existing exciter a difficult job.

to build a closed-loop ssb system with rf clipping. The output will be at the original audio frequency with instantaneous amplitude limiting, just as in a speech clipper, but without any new harmonic components. Therefore, it is suitable for the speech input of a ssb transmitter (and will give superior performance when used with a-m and fm equipment).

Fig. 5 shows the block diagram of such

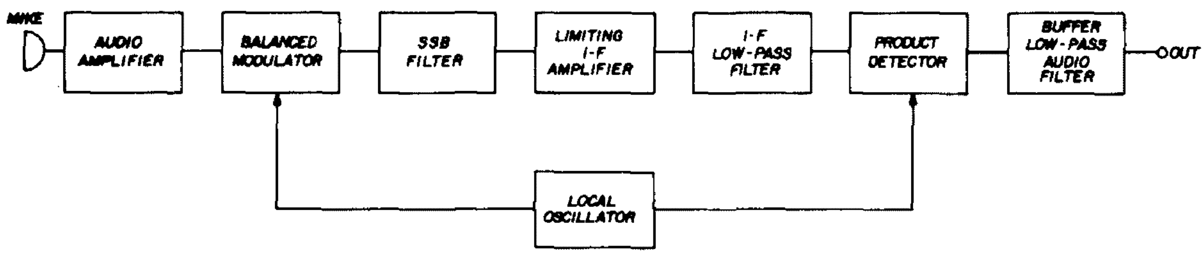


fig. 5. Block diagram of the Comdel CSP-11, an audio accessory unit with rf clipping.

Rf clipping is at least as good as speech clipping with a-m with the advantage that distortion is small compared with the older method. A 10-dB gain in average power can be expected with 15-dB clipping. The actual gains may be lower or higher depending on the severity and nature of the difficulty of the communication path. In 1961, Paul Day, W1PYM, and I demonstrated and recorded the effectiveness of rf clipping by simulating a poor ssb link. An unprocessed transmission was adjusted to be on the threshold of readability, say R2. A transmission with the identical peak output power and incorporating speech clipping produced no improvement. In contrast, a transmission using rf clipping produced a readability of 4 to 5. In fact, the peak output could be reduced considerably below that of the unprocessed transmission before intelligibility was affected.

compatible audio accessory unit

Rf clipping is probably the most effective way to increase the talk power of a ssb transmitter. Since modifying an exciter to incorporate the feature is difficult at best, a way around the problem is

an accessory unit, the Comdei CSP-11. Note that there can be no frequency error since the same oscillator is used for modulation and demodulation. In this unit peaks are limited or clipped instantaneously, there is no discharge time constant, and harmonic distortion is absent.

While experimenting with rf clipping the scheme shown in **fig. 5** was realized and tested. There was no audibly discernible difference between the performance with rf clipping and that obtained with the audio accessory unit. (In theory, the latter should produce slightly higher IM distortion.) The audio was demonstrated to a member of a well known audio consultant firm; his first reaction was disbelief that he was listening to a clipping system! His own tests with the processed audio (there was no rf link involved) showed that:

1. When mixed with noise, the new method is slightly superior to normal speech clipping; 15 dB clipping produced 10 dB intelligibility gain with no apparent distortion (in noise).

2. There was no loss of intelligibility whatever "in the quiet" up to 24 dB clipping. In regular speech clipping

systems where there is some loss because of distortion.

Many years of experience have convinced me of the value of rf clipping in the simulated form. Unless specifically requested for demonstration purposes I never switch the device out of circuit and often receive unsolicited gratifying reports.

conclusion

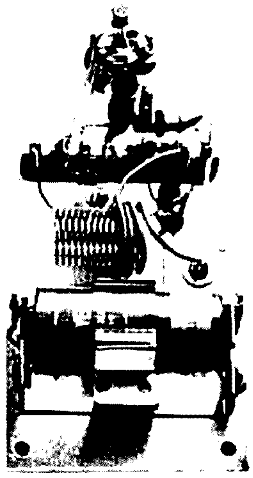
Speech clipping in its old form, or any device which causes severe distortion of the audio signal, should not be used with ssb transmitters. Properly designed alc systems and volume compressors can be used to prevent flat-topping and give moderate gain in talk power. With regard to talk power, rf clipping and its audio derivative will give the best results, but the devices are rather costly.

Like most things in life, processing the ssb signal can be overdone and the most effective devices misused. For instance, moderation is in order when the microphone is in a noisy location. Most of us are familiar with the "aeronautical mobile" effect where the noise power is as high as that of the intelligence whenever the operator stops speaking. While the signal-to-noise ratio is not actually degraded the effect on the listener is one of annoyance. It is well to ensure that the PEP output of a transmitter with speech is at least 15 dB – preferably 20 dB – *above the noise* PEP.

I hope that I have managed to convince you that speech clipping is *not* for the ssb station. There are other devices and techniques available, though they cost a bit more. When viewed against the total amateur station investment the additional expense is really quite reasonable.

For those of you who want the ultimate in performance (and cost) a local long-time-constant agc loop ahead of the rf clipper will ensure that a fixed amount of clipping, say 18 dB, can never be exceeded. A properly adjusted volume compressor can be made to serve the same purpose.

ham radio



field-effect transistor transmitters

Low-power transmitters for two and ten meters that use field-effect transistors in every stage

Albert D. Helfrick, K2BLA, Paramus, New Jersey 07652

The field-effect transistor has recently been finding its place in radio communications equipment. Although its performance as an rf amplifier and mixer is well known the fet's merit in other functions seems to have gone unnoticed. The fet can function well in dc amplifiers, audio amplifiers, switching circuits, oscillators, multipliers and phase modulators. This article presents a low-powered vhf transmitter that uses field-effect transistors in every stage.

Although the fet deserves consideration when designing vhf transmitters it will not quell the nightmares of the solid-state vhf transmitter designer. It has its advantages: inexpensive, simple class-C biasing, low feedback capacitance, good efficiency with a 12-volt supply and relatively high power gain. The fet also has its disadvantage – low power dissipation. Commercially available field-effect transistors were developed primarily for small-signal use and typical power dissipation ratings are on the order of 0.4 watt. Since these ratings are based on no external heat sinking it is possible to

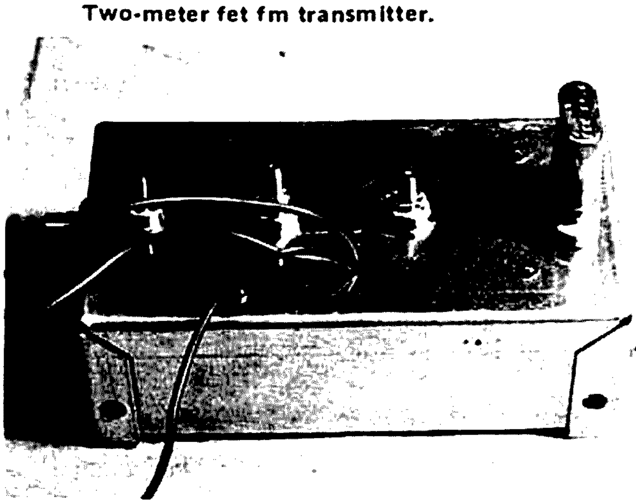
decrease the case temperature and operate the fet beyond published ratings. However, for low-powered transmitters or low-power stages of higher powered transmitters, field-effect transistors can be used as is.

I have tried several experiments with transmitting circuits using fets, including fundamental and overtone crystal oscillators, frequency multipliers, rf amplifiers, and phase modulators. The two transmitters described here use experimental circuits. Both n-channel junction fets and n-channel depletion-mode mosfets were used. Mosfets and jfets are usually interchangeable but in most cases the simplicity of gate-leak bias for class-C circuits favors the jfet. The same results can be obtained by adding a diode from gate to source in the mosfet. In most cases performance is sufficient and the use of a mosfet doesn't warrant the additional component.

two-meter fm transmitter

The two-meter fm transmitter in **fig. 1** uses five junction field-effect transistors. It is intended to be used with a portable vhf receiver as an fm walkie-talkie. Using a Heathkit GR88 vhf monitor receiver (tuned below its normal range) a range of over one-half mile was obtained between two walkie-talkies. An additional amplifier, tube or transistor, could be added for additional power for more serious work.

Two-meter fet fm transmitter.

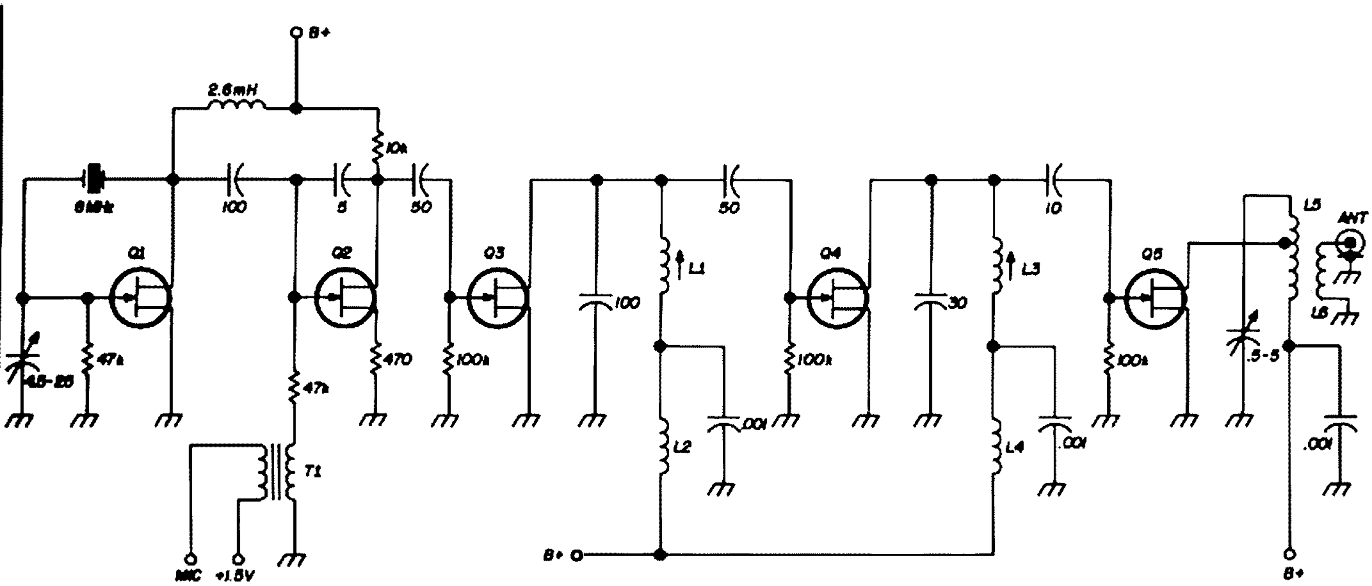


The cost of the fet transmitter is quite low. Homemade coils and chokes were used along with bargain-variety transistors. Devices were selected for optimum

Bottom view of the two-meter fet transmitter shows component layout.

performance in the 72-MHz tripler and the 144-MHz doubler stages. Excluding the crystal the total cost was below $10.00.

In this transmitter Q1 operates as an 8-MHz Pierce crystal oscillator which drives the phase modulator. The phase modulator was designed after a circuit used in an antique Link high-band mobile transmitter. The Link modulator provided 20-kHz deviation at 144 MHz using a 3.0-MHz crystal. I had no difficulty obtaining 5-kHz deviation from an 8-MHz crystal in the fet version. The modulator drives a single tripler tuned to 24 MHz,



L1 9 turns no. 28, closewound on a ¼" slug-tuned coil form

L2 0.75 μH (J. W. Miller 4651)

L3 3½ turns no. 20, closewound on a ½" slug-tuned coil form

L4 10 turns no. 20, ¼" diameter

L5 3 turns no. 18, ½" diameter, 6 turns per inch, tapped at 2 turns (Air-Dux 406T)

L6 1 turn hook-up wire around cold end of L5

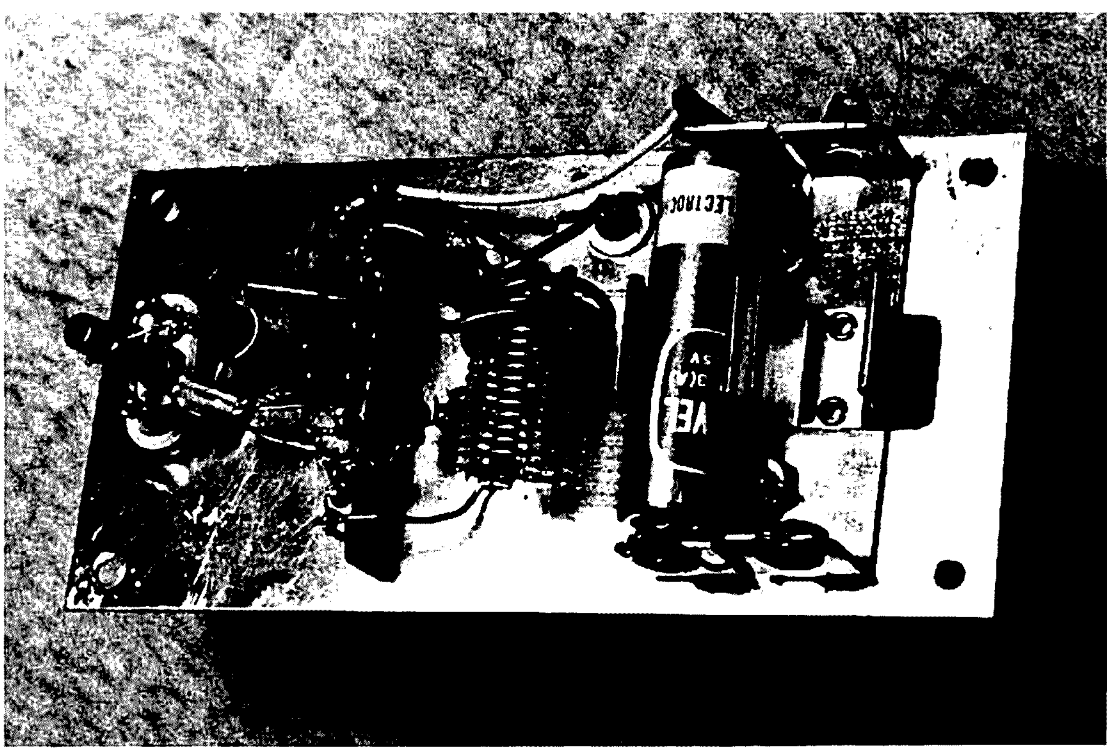
T1 Carbon-mic to grid transformer

fig. 1. Schematic diagram of the two-meter fet phase-modulated fm transmitter.

which drives another single-tuned tripler and a doubler. The efficiency of the higher frequency multipliers decreases with increasing frequency, and it was necessary to select fets for optimum performance. Since the single-tuned multipliers offer little rejection of unwanted harmonics it is desireable to double tune the 144-MHz doubler if the unit is to drive a higher powered amplifier.

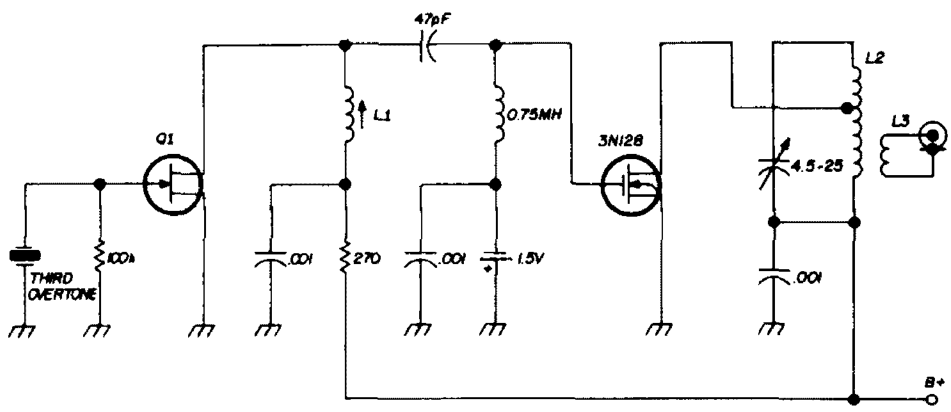
The transmitter was originally designed for a 9-volt power supply to conform with the monitor receiver. However, fet multipliers, like their vacuum-tube counterparts, require the highest supply voltage permissable for maximum

Ten-meter transmitter features a mosfet in the power-amplifier stage. The battery provides bias only, and could be replaced with a miniature mercury type.



efficiency. The fets used in the transmitter had a maximum V_{ds} of 20 volts and, as expected, best operation was obtained with a supply voltage of 20 volts. Two

The final amplifier in the ten-meter transmitter uses a single 3N128 mosfet with a 1.5-V dry cell for fixed bias. Since the drain on the bias battery is essentially



L1 12 turns no. 36, closewound on ¼" slug-tuned coil form

L2 13 turns no. 18, tapped at 8 turns (Air-Dux 408T)

L3 3 turns hookup wire around cold end of L2

fig. 2. Ten-meter transmitter using a mosfet power-amplifier stage. Power output is about 200mW.

9-volt batteries in series were used for portable operation; 9 volts for the receiver and 18 for the transmitter. A small dry cell was used for the carbon microphone.

ten-meter transmitter

The transmitter in **fig. 2** is capable of 200 mW output into a 50-ohm load at 29 MHz with an overall *efficiency* of 50%.* The transmitter is a two-stage affair consisting of a jfet overtone oscillator and a class-C mosfet amplifier.

I have had great success with fet crystal and self-excited oscillators. Not one fet oscillator failed to oscillate, and all the experimental circuits exhibited excellent frequency stability. In the tuned drain-oscillator in the ten-meter transmitter it was possible to operate fundamental crystals in the third overtone, and to operate overtone crystals in the fundamental simply by changing the tank resonance.

*Overall efficiency is the total transmitter power input divided into the output power. This includes the power for the oscillator.

zero, a small 1.4-volt button-type mercury cell could be used since its shelf life is several years.

Two fets were tried in parallel with a small increase in power output but a reduction in efficiency. Previous experiments with parallel mosfet amplifiers at 50 MHz showed some hope, but most circuits, when pushed beyond one-watt output, resulted in burned out mosfets. Because of the difference in mosfet characteristics, one transistor does all the work while additional parallel devices only decrease efficiency. The single 3N128 required no neutralization and provided almost 200 mW output with a 12-volt supply.

The purpose of the ten-meter transmitter is the same as that of the two-meter fm unit. Many inexpensive 30- to 50-MHz monitor receivers can easily be tuned to 29 MHz, converted to a-m, and used with the fet transmitter. From my experience with 100-mW citizens-band transceivers on 10 meters, the range of such a combination should be quite respectable.

ham radio

improving the Motorola P-33 series

Add these modifications and you'll have a truly high-performance 2-meter rig

Richard Zach, WB2AEB, 22 Pike Place, RFD-4, Mahopac, New York 10541

Some authors have referred to fm equipment as the "new surplus."[1] If you were to choose the ARC-5 equivalent of the new surplus, it would no doubt be the Motorola P-33. This is a 5-watt unit with a partially or fully transistorized receiver (P-33A or B respectively).

The P-33 transmitter uses quick-heating tubes. Power can be supplied by nicad batteries, dry cells, or a 6/12-volt power supply. The P-33 is readily available, usually at less than $100.00. In this part of the country it seems as though every ham on fm uses one.

The P-33 has two relatives. One is the H-23A (or B), which is a one-watt handy-talky. Its big brother, the D-33A (or B) Dispatcher, is a 10-watt motorcycle unit. These rigs have essentially the same transmitter and receiver strips, so the following modifications are applicable to all three models. Included are:

1. Changes for receiver 2-meter operation and an fet front end.
2. Changes to receiver and transmitter to increase bandwidth.
3. Improvements to the nicad-battery supply in the P-33BAM equipment.

receiver front end

The fet addition to the receiver was introduced to me by Walt Fairbrother, W1RYL. In one of my earlier articles, I vaguely mentioned this modification, and both Walt and I were deluged with mail inquiries. Since then I've made a few minor improvements.

The P-33 series was originally in two forms, low high-band (136-150 MHz) and high high-band (150-174 MHz). Most available units are of the high-frequency version. The following modification will improve sensitivity and also put the unit in the two-meter band. For exact pin locations and tune-up instructions, consult the manual for your unit. The receiver modifications are shown in fig. 1.

1. Remove the input coax cable at L1 pin 4.

2. Remove the jumper between L1 pin 2 and L2 pin 3.

3. Carefully remove the shield can from L1 by unsoldering the two lugs that hold it to the PC board.

4. Remove C1 (9 pF), C2 (47 pF), and C3 (.51 pF).

5. Disconnect the grounded end of L1 from the board.

6. Carefully drill out the pin-1 hole where the grounded end of L1 was connected. Start with a no. 60 drill to make a pilot hole. Enlarge the hole further, using a no. 51 drill. Be careful not to allow metal chips to fall into the transmitter section.

7. Using a razor blade, cut out and peel off the conductive material on the bottom of L1 (see **fig. 2**). Do this to both sides of the PC board. This will insulate pin 1.

8. Using a no. 27 drill, slightly countersink both sides of the hole.

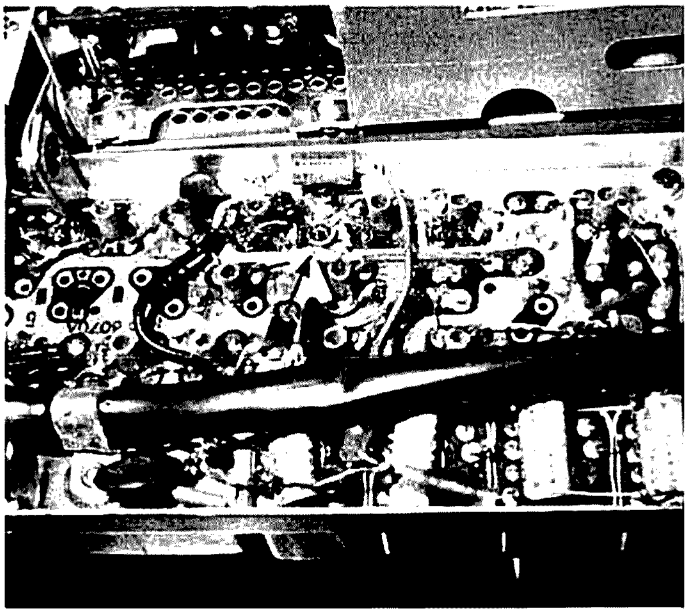
9. Insert the new C2 (56 pF) into holes 1 and 4 of L1.

10. Insert the cold end of L1 coil into hole 1.

11. Insert the new C1 (12 pF) into holes 3 and 4 of L1.

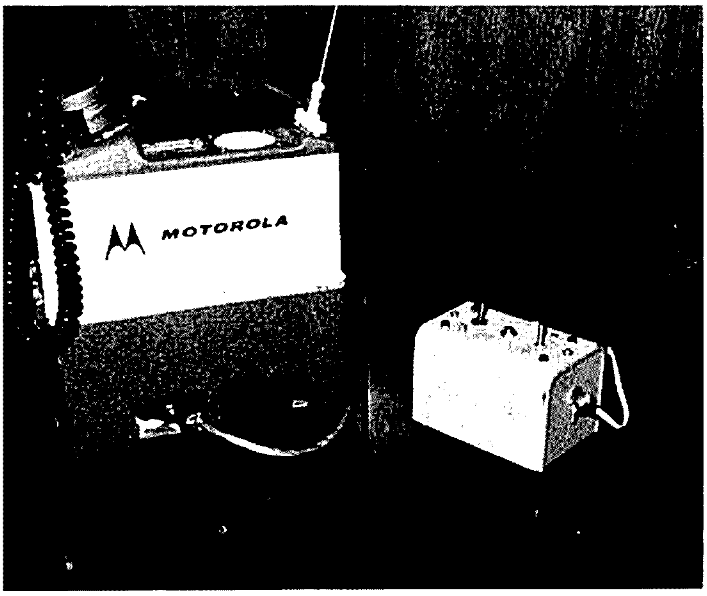
12. Solder L1 pins 1, 3, and 4.

13. Drill a pilot hole for the 3N128 between L1 and L2.



Bottom of receiver board showing the fet leads and other wiring changes. C54 can be seen near cable clamp (see receiver widebanding).

Motorola P-33 portable with homebrew battery charger. Using the jack on the side of the unit, it can be powered by an external 12 Vdc source.



14. Using successively larger drills, enlarge the hole so that the transistor case fits snugly. You should end up using a no. 12 drill.

15. Remove the can from L2 as in step 3 and remove C4 (6.6 pF). Replace it with the 9 pF capacitor (C1) removed in step 4.

16. Remove the can from L3 and remove C6 (5 pF).

17. Replace this capacitor with the 6.6 pF unit (C4).

18. Solder the shield cans onto L1, L2, and L3.

19. Install a .001 μF capacitor between L1 pin 1 and ground.

20. Install a 200-ohm, 1/8-watt resistor between L1 pin 1 and the (now vacant) pin 2. (Pin 2 will become a tie-point.)

21. Connect a small .001 μF capacitor between L1 pin 2 and ground.

22. Insert the transistor into the hole between L1 and L2. The leads should face away from the case. *Do not* remove the static discharge wire from the transistor at this time.

23. Ground pin 4 of the transistor.

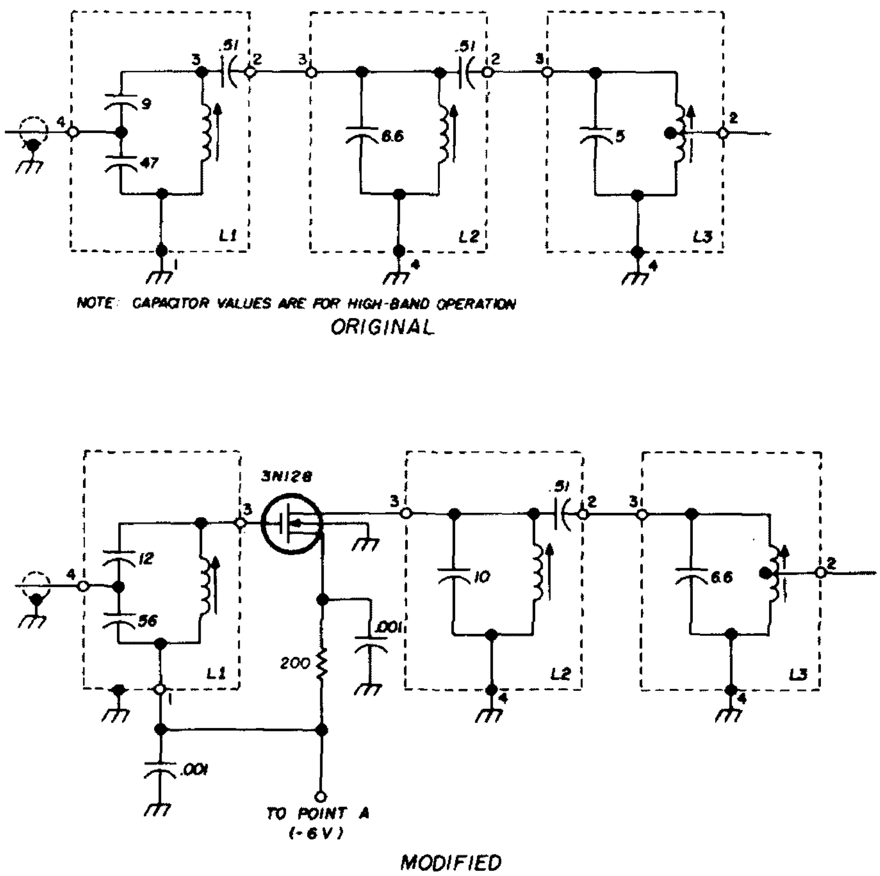


fig. 1. P-33 receiver front-end modifications per W1RYL. Before and after circuits are shown in A and B.

table 1. Accessories for the P-33 available from Motorola.*

Item	Part No.	Price
NB Permakay filter	NFN 6000 AS	$21.00
WB Permakay filter	NFN 6000 AW	21.00
Tuning tool	66A847036	1.15
Tube	type 6397	7.84
Manual for:	68P81005A40-E	1.50
P-33BAM		
P-33BAC		
H-23BAM		
H-23BAC		

*1875 Greenleaf Avenue, Elk Grove Village, Illinois 60007.

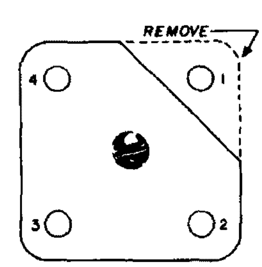


fig. 2. Conductive material must be removed as shown from both sides of the PC board under coil L1 for receiver front-end mods.

24. Connect lead 3 of the transistor to L1 pin 3.

25. Connect transistor lead 2 to L1 pin 2.

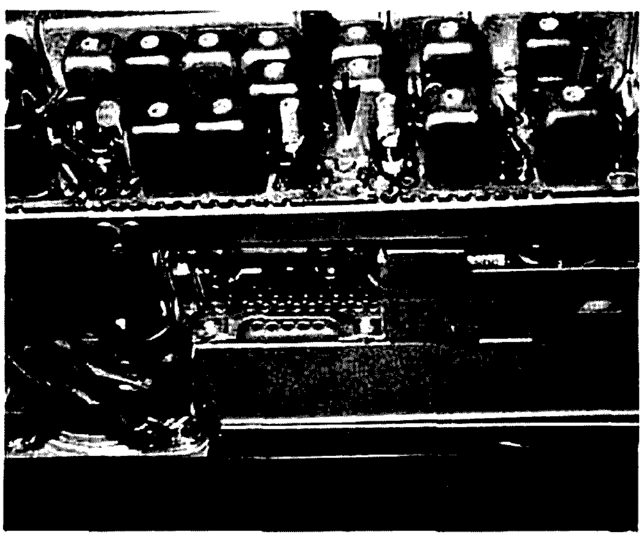
26. Connect transistor lead 1 to L2 pin 3.

27. Run a wire from L1 pin 1 to point A (see fig. 1). (Point A is between L11 and L9.)

28. Connect the inside conductor of the antenna coax cable to L1 pin 4 and ground the shield.

29. Remove the static-discharge wire from the transistor.

30. Retune the front end as described in the manual.



Top of receiver board showing L1 and L2 without cans. The top of the fet can be seen between the coils just above the screw.

increasing bandwidth

If you don't know whether your receiver is wide or narrow band, look for a long, thin Permakay filter in the receiver section. If the receiver is a wideband unit, the filter will be numbered NFN 6000 AW. Narrowband receivers will have the number NFN 6000 AS.

If narrowband operation is used in your area, you'll be delighted to know that most P-33s are narrowband. However, if your area uses the more common wideband deviation, start warming up the soldering iron.

1. Remove C54. This is a 2000-pF capacitor at the output of the Permakay filter. It's located very close to the clamp that secures the control cable.

2. Connect a small 56k resistor between L30 discriminator can pin 10 and pin 6.

3. Connect another small 56k resistor between pins 8 and 9 of the same can, L30.

4. This step is optional (and expensive), but the change greatly improves receiver audio quality. Remove the narrowband filter (NFN 6000 AS) and replace it with the wideband version (NFN 6000 AW). See table 1.

To increase transmitter bandwidth, simply adjust the deviation pot (labeled IDC) for the appropriate level. This control is located near a small transformer (T1) on the transmitter board.

table 2. Capacitors to be added to the P-33 transmitter. (All are 500 Vdcw.)

Capacitance (pF)	Connections
8.2	V3 pin 1 to ground
8.2	V4 pin 4 to ground
2.0	V5 pin 1 to ground
1.5 20%	across L8
1.5 20%	from one side of L9 to L9 center tap
1.5 20%	from other side of L9 to L9 center tap
1.5 20%	across L10

transmitter improvements

Getting the transmitter to operate on two meters is no major problem. If you want optimum performance from your unit, pad the transmitter with the components shown in table 2.

After you tune up the transmitter, check it for rf output. See table 3 for the power output to be expected from your unit. If output is low, very carefully retune the transmitter and power amplifier. Also adjust R35 for 28 mA at JU-1.

table 3. Power output to be expected from H-23 and P-33 units with various power-supply conditions.

H-23 Series
- **1.0 watt at full battery voltage (135 Vdc)**
- **0.8 watt at nominal battery voltage (120 Vdc)**

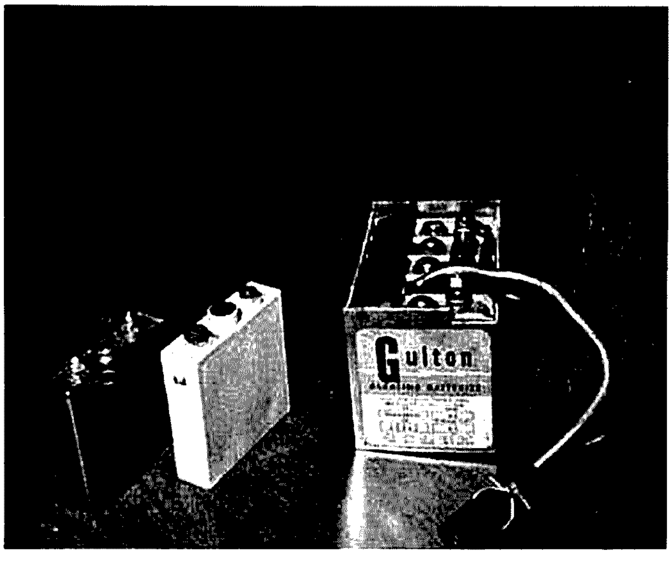
P-33 Series
- **3 watts at 162 volts (dry cell nominal)**
- **4 watts at 180 volts (dry cell maximum)**
- **5 watts at 240 volts (nicad batteries)**

multiplier and driver checks

It's very difficult to find a tube checker to test the multiplier and driver chain, so check the 2E24 in the power-amplifier cage, even though this tube seemingly never needs changing. After you've made further steps, such as voltage checks, you can feel confident that the third doubler and drivers are at fault. This is a common problem with P-33s. All these tubes are 6397s, and they'll set you back a fat $7.84 each.

Why did they go bad? There are a couple of possible reasons. The 6397 has a quick-heating filament. After awhile, a filament may not heat as quickly as it should. While transmitting, the driver tubes may not be getting any excitation, and they'll blow. Thus, all three usually go in "domino" fashion.

Nicad batteries in P-33BAM, H-23BAM series. From left, metal vented nicad cell; plastic-coated non-vented cell (nicad); vented nicad battery.



The second cause of tube failure is preventable. Some hams operate the P-33 directly from the car's ignition system. If a large enough voltage spike occurs, the 6397s will be zapped. Many fm'ers run their units like this with no trouble; then again, some aren't so lucky.

power-supply improvements

In equipment such as the P-33BAM, the "M" denotes a nicad-battery supply. Likewise, a "C" denotes a dry-battery pack in a P-33BAC.

There's not much to say about the dry-battery pack, because it's merely a box. The nicad supply, however, is a 6/12-volt dc-to-dc converter. Nothing seems to go wrong with the inverter, but every P-33 owner should become an expert on the care of the nicad batteries. After all, a new set of nicads would cost you only $128.50! Used nicads can be obtained for about $15, though, thanks to the surplus emporiums.

It's impossible to give a complete discussion on nicads here, but the following is advisable. If the nicad pack is vented, each cell will have a rubber gasket and a screw at the top. Sometimes the electrolyte leaks out of the vent and solidifies. Simply clean off the entire battery with clear water and a toothbrush. Make sure the vents are screwed shut when cleaning. If electrolyte is needed, place approximately two potassium hydroxide (KOH) pellets into the required amount of water for each cell. Do not saturate the electrolyte. Mix the solution well before putting it into the cell. (This procedure isn't necessary for nonvented cells.)

Incidentally, some of the batteries made by Gulton (for Motorola) are labeled "Alkaline Batteries." These are actually nicads. The batteries made by NIFE are plainly labeled nicads.

charging nicads

Measure the voltage of each cell. A fully charged cell should show 1.25 volts under load. If the cell shows zero volts, check it to determine if it's shorted. If so, the shorted cell should be replaced.

Each battery consists of five cells, and there are two batteries in the P-33. The H-23 has one battery, and the D-33 has none.

If you don't have a charger, the circuit shown in fig. 3 will do the job nicely. An advantage of this circuit is that it will charge the batteries while they're installed in the P-33. You can also simultaneously charge batteries and operate the rig, providing you have a *regulated* 12-Vdc source.

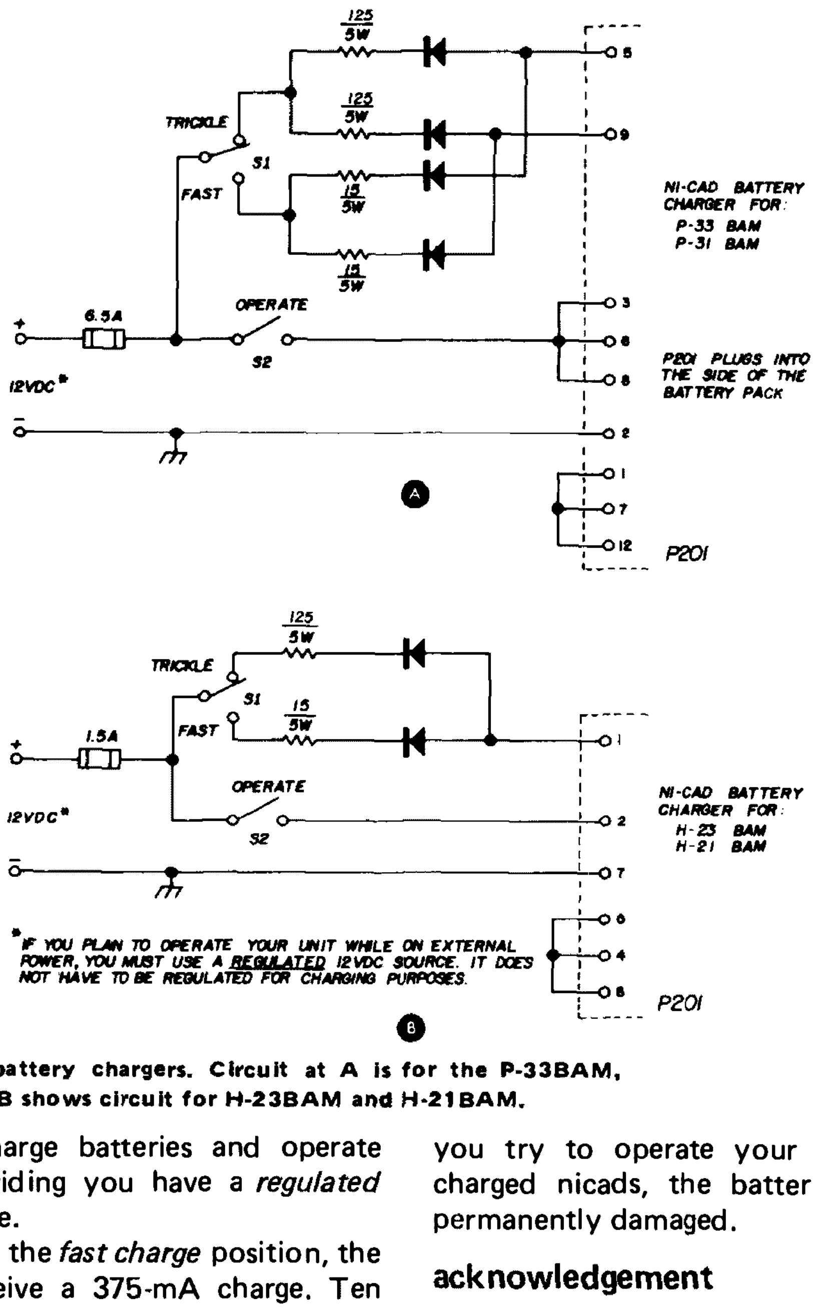


fig. 3. Nicad battery chargers. Circuit at A is for the P-33BAM, P-31BAM, etc. B shows circuit for H-23BAM and H-21BAM.

With S1 in the *fast charge* position, the batteries receive a 375-mA charge. Ten hours are required to charge the batteries at a 400-mA rate; however, when S1 is in the *fast charge* position, the circuit charges at 375 mA as a safety precaution. Thus if the battery is fully discharged it should be charged for 10 hours; if half discharged, it should be charged for 5 hours, etc. A word of caution is an order. If S1 remains in the *fast charge* position too long, the nicads may explode.

When the batteries are fully charged, S1 may be placed to *trickle charge.* This will ensure fully charged batteries when you need them. The batteries can be trickle-charged continuously with no harmful effects. Discharged nicads should be recharged immediately after use. If you try to operate your P-33 on discharged nicads, the batteries might be permanently damaged.

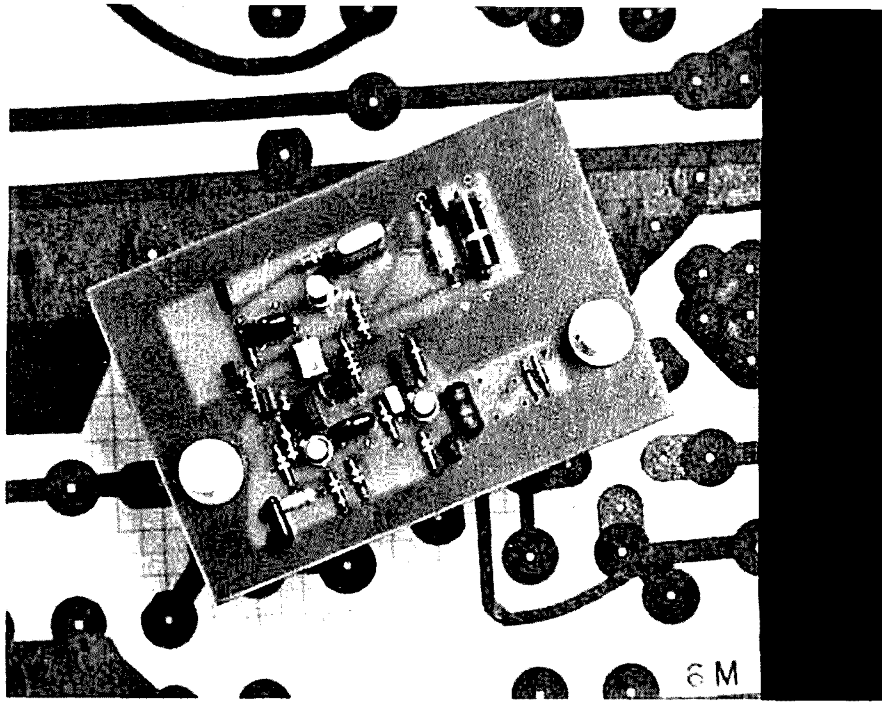
acknowledgement

I'd like to express my thanks to WA2HXY and W1RYL for their assistance in the preparation of this article.

reference

1. L. Cobb, W6TEE, and J. O'Brien, W6GDO, "Amateur Fm and Repeaters," *QST,* October, 1969, pp. 11-13.

ham radio



deluxe mosfet converters for six and two meters

High-performance vhf converters featuring gate-protected mosfet devices and simple construction

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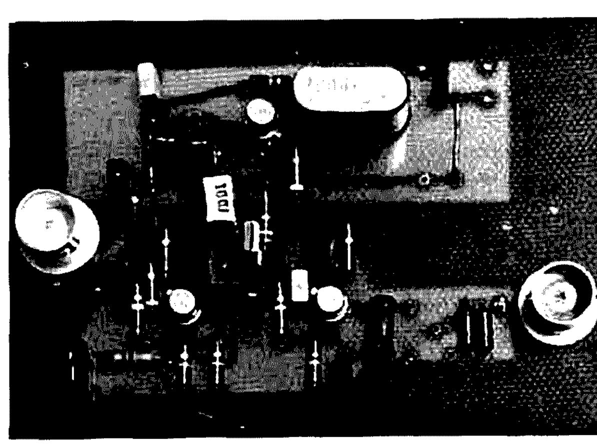
Since 1968 when the first mosfet amateur receiving converters were described in *ham radio,*[1] many hams have had the opportunity to prove their worth. I know of 20 such converters operating in the Philadelphia area alone. While the original design was easy to construct and produced troublefree operation, a redesign using a printed-circuit board and the latest transistors further simplifies the project. By combining the experience of many users it is possible to direct the builder in customizing his own system.

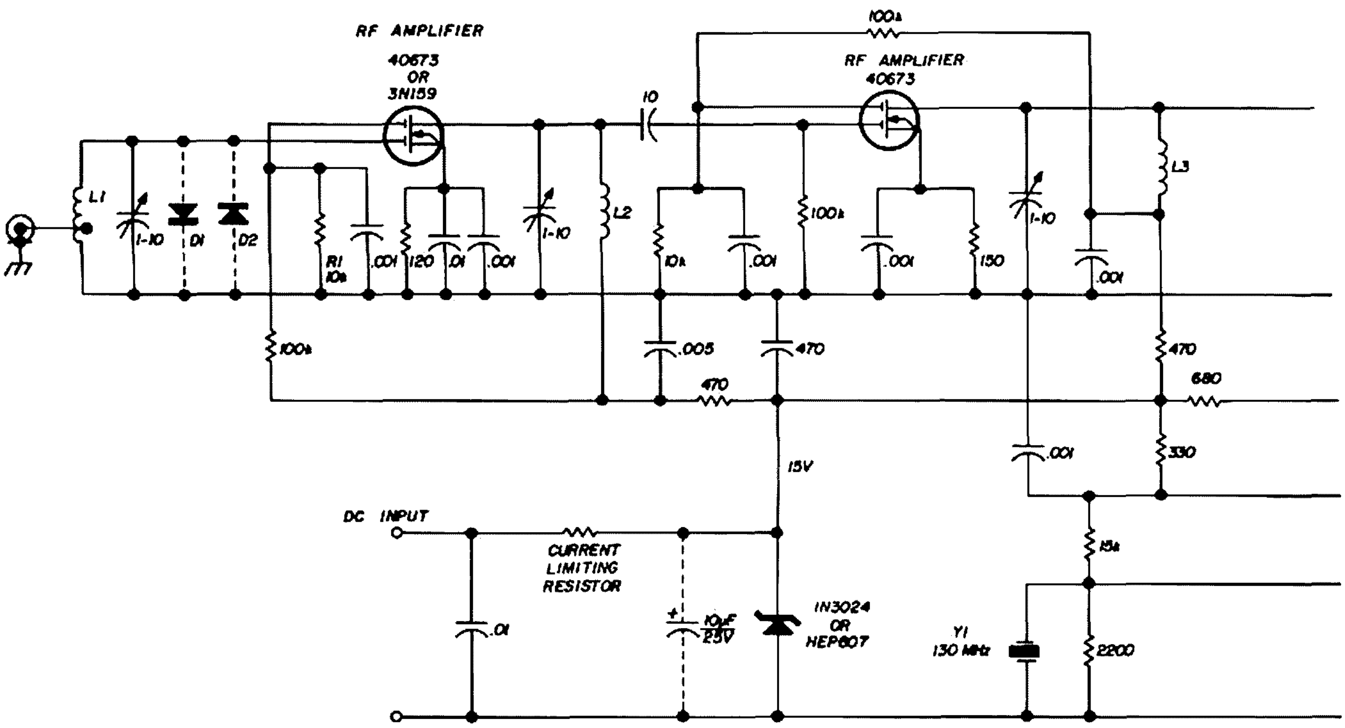
Top view of the six-meter converter. No zener regulation was used in this unit. I-f output is at 14 MHz.



choosing transistors

With the introduction of the gate-protected mosfet, most of the former problems with premature burnout are eliminated. At the time of this writing, there are three such transistors, the RCA 40673, 3N187 and 3N200. The 3N187 and 3N200 are basically the same as the 40673 but feature more tightly controlled operating characteristics. A list of protected and unprotected mosfets suitable for vhf use is given in table 1.

One disadvantage of the 40673, which may be overcome by using other mosfets, is the wide variation in characteristics that is possible. By selecting a 3N159 for the front end, the lowest noise figure is possible. Next in order of preference is the 3N140. This is also a good choice for the rf stage of the 6-meter unit and the second rf stage of the 2-meter converter. Performance comparable to the 3N140 is obtainable from Motorola's MFE3007. An RCA 40603 or Motorola MFE3006 would also be suitable as rf amplifier at 50 MHz. In the mixer — the least critical stage — lower gain transistors such as the 3N142 or MFE3008 are useful.

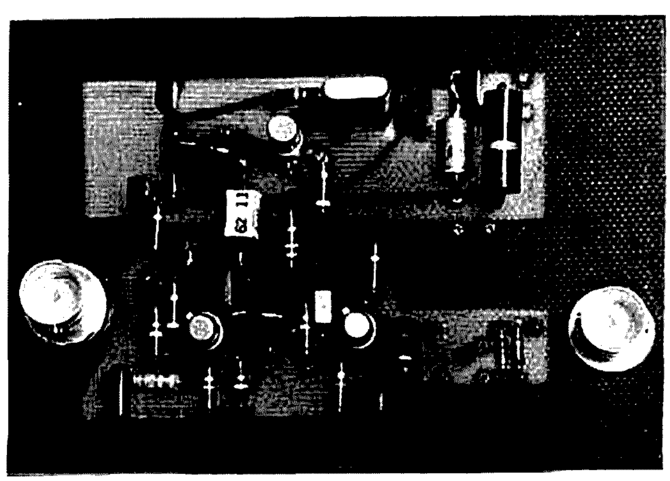


mosfet handling precautions

Since electrostatic charges can destroy the gate insulation on mos field-effect transistors, the devices must be handled with a certain amount of care. The following handling procedures are recommended:

1. Do not remove the external shorting

This six-meter mosfet converter has a 28-MHz i-f output. Note the zener regulator and large voltage-dropping resistor. The crystal is a small HC-18 type.



wire until all components are wired into the circuit.

2. Use a soldering iron with a grounded tip (3 wire).

The RCA 40673, 3N187 or 3N200 do not have an external shorting spring because they are internally protected against low energy charges. However, precautions 2 and 4 still apply. Mosfets

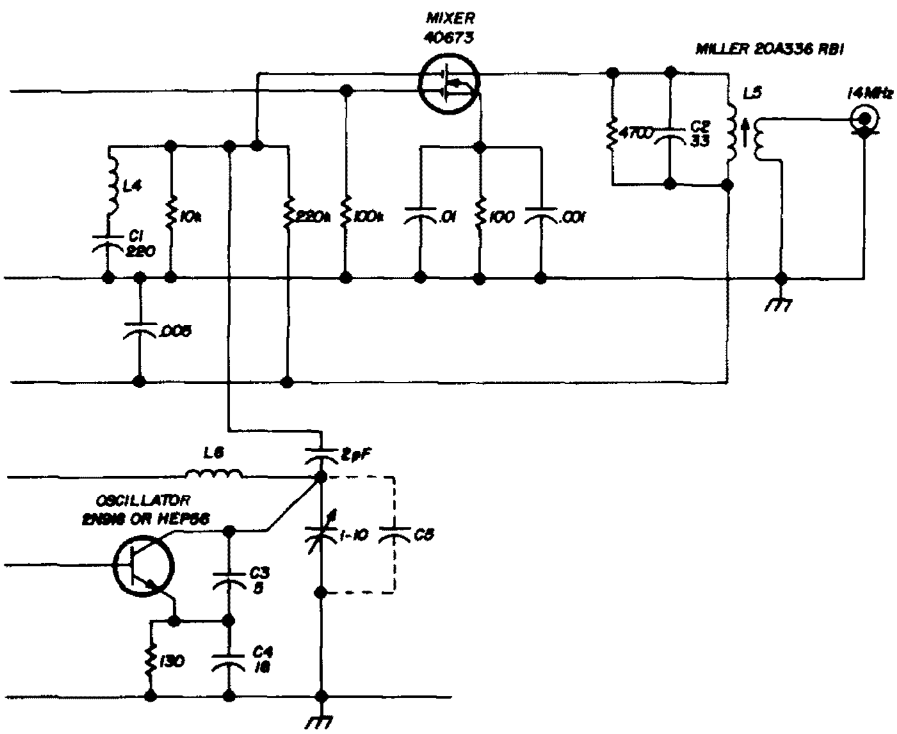


fig. 1. Circuit diagram for the deluxe mosfet two-meter converter. R1, for best noise figure, should be 33k; for maximum gain, 47k; for minimum cross modulation, 10k. Components given in parts list are for 28-MHz output.

L1 5 turns no. 16, ¼" diameter, tapped 1¼ turn from ground, air wound

L2 4 turns no. 22, ¼" diameter, air wound

L3 3 turns no. 22, ¼" diameter, air wound

L4 0.68 µH (J. W. Miller 4590)

L5 2.38 — 3.96 µH (J. W. Miller 20A336RBI) with 3-turns no. 26 output link

L6 5 turns no. 22, ¼" diameter, air wound

3. If any components must be added or removed, short the mosfet leads before proceeding.

4. If possible, isolate test equipment probes with a capacitor; some test equipment, usually old, has high voltages present which will destroy the transistors.

from Motorola such as the MFE 3007 are built differently and have higher breakdown voltages – one experienced user told me he experiences much less trouble with these. My own personal problems with mosfets have been few, but good clean living helps.

Both converters shown in this article are basically the same as the original

table 1. Operating characteristics of mosfets suitable for vhf converters.

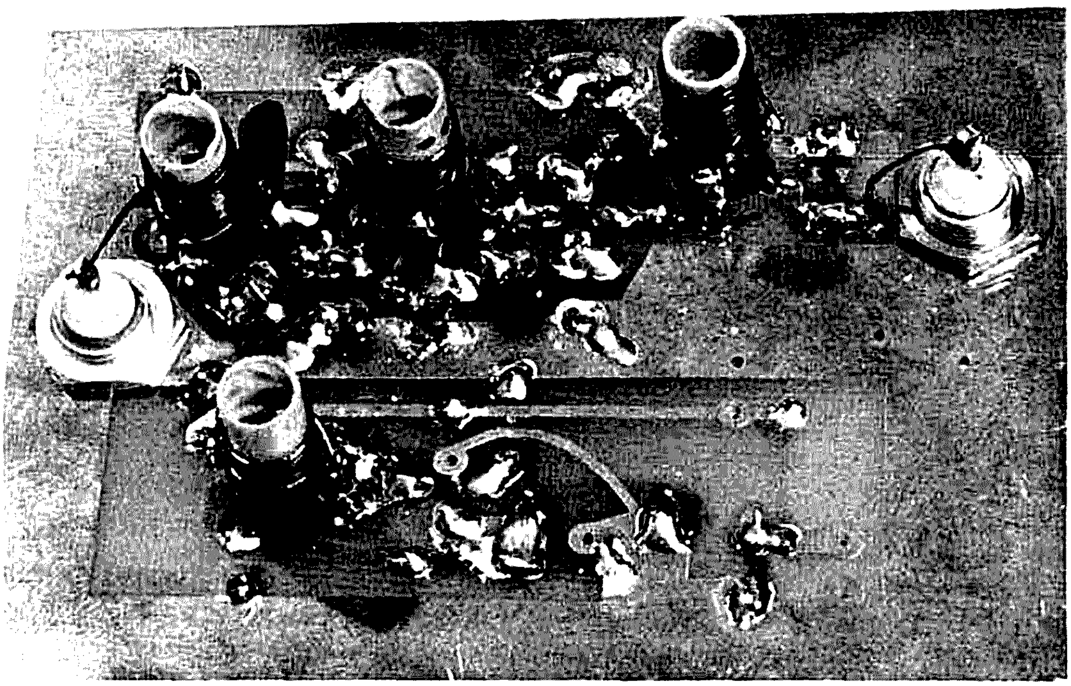
protected type	40673	3N187	3N200		
closest unprotected equivalent	3N140	3N140	3N159 or TA7153	3N159	TA7153*
maximum operating frequency	400 MHz	300 MHz	500 MHz	300 MHz	500 MHz
typical power gain at 200 MHz	18 dB	18 dB	19 dB	18 dB	na
typical noise figure at 200 MHz	3.0 dB	3.5 dB	3.0 dB	2.5 dB	na
maximum noise figure at 200 MHz	6.0 dB	4.5 dB	na	3.5 dB	na
typical power gain at 400 MHz	na	na	12.5 dB	na	14 dB
typical noise figure at 400 MHz	na	na	4.5 dB	na	4.5 dB
disadvantages	wide tolerances	—	—	unprotected gates	unprotected gates

na indicates information is not available
*not distributed through normal channels

design,[1],[2] although dual-gate mosfets are used throughout. One of the big advantages of the dual-gate configuration is that the converter may be optimized for gain, noise figure *or* cross-modulation resistance by changing one resistor, R1. This resistor, from gate 2 of the first rf stage to ground, should be 10k for lowest cross modulation, 33k for best noise figure, and 47k for higher gain. There is little advantage in making R1 larger than 47k because you can exceed the input rating of the transistors. The values of R1 shown here are experimental, and the operating characteristics will vary somewhat from one device to another.

Experiments with a hot-carrier-diode balanced mixer instead of the dual-gate mosfet were interesting. I found that unless the diodes were matched, there was no clear reduction in cross modulation over the mosfet stage.* In the presence of extremely strong signals, the two-rf-stage converter will have reduced gain – possibly to the point where weak signal reception is impractical. A single rf stage almost completely eliminates the effect.

The cost of unmatched hot-carrier diodes and their associated circuitry is four to five times more than the mosfet equivalent, and matched diode assemblies are even more expensive. From my experience, I feel that hot-carrier diodes should be reserved for higher frequency converters and other applications.

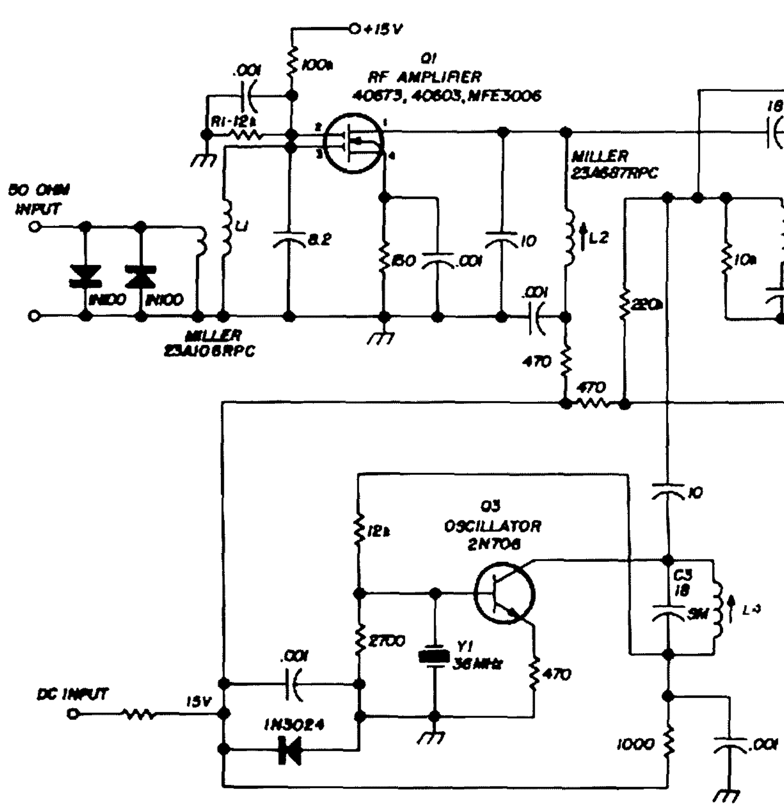


Bottom view of the six-meter converter with 14-MHz output. Coils have been mounted on this side to conserve space.

The oscillator of the two-meter converter has been upgraded by using a crystal at the injection frequency. This has the advantage of reducing the number of images. The higher cost of the crystal is compensated by reducing the number of components in the circuit.

The back-to-back diodes at the input are a very controversial subject. The amateur

*The experiment was conducted on two meters, with a 3N141 mosfet mixer, and a hot-carrier-diode mixer using Hewlett-Packard HP2900 diodes.



L1 0.76 — 1.25 μH (J. W. Miller 23A106RPC) with with 3-turn input link

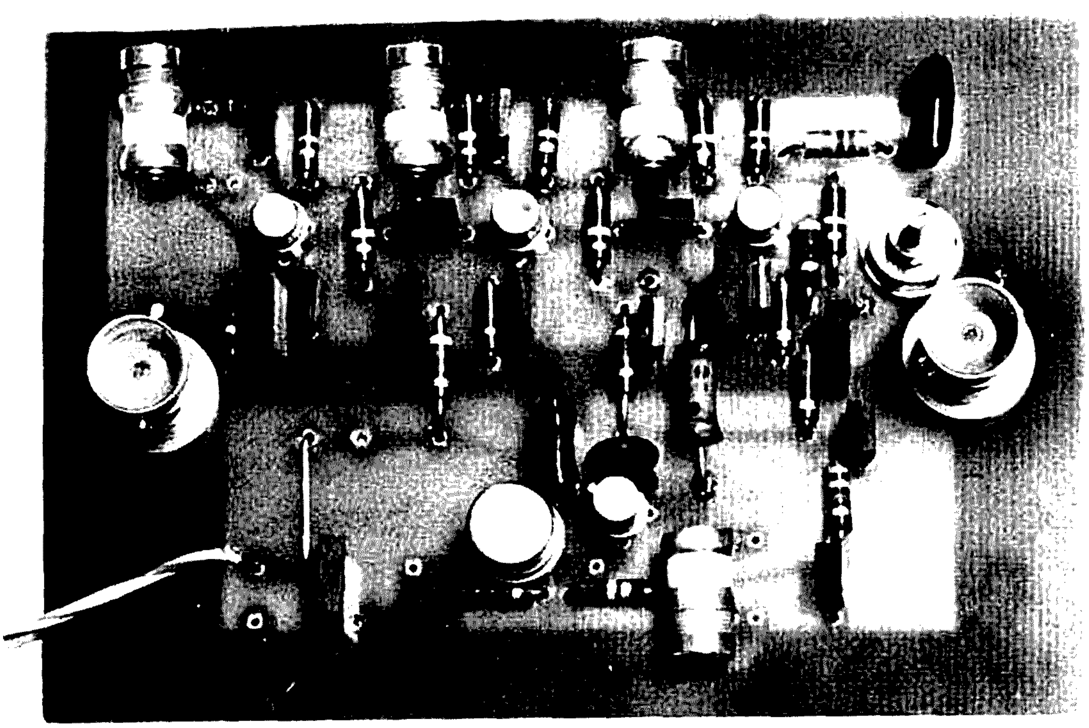
L2 0.54 — 0.85 μH (J. W. Miller 23A687RPC)

L3 1.65 — 2.75 μH (J. W. Miller 23A226RPC) with 3-turns no. 26 output link

L4 1.20 — 1.87 μH (J. W. Miller 23A156RPC)

fig. 2. Circuit diagram for the deluxe mosfet six-meter converter. Components with an asterisk are for operation from higher voltages (see text). Components in parts list are for 28-MHz output.

Top view of the two-meter converter. A special crystal is used in this model, but the more common HC-18 will fit equally well.



with a kilowatt (on any band) is going to need some protection, and the built-in diodes on the 40673 may not be enough. While 1N100s or 1N916s may protect the transistors, they may cause TVI during transmission and severe cross modulation when receiving if a high-power transmitter is nearby. My answer is to use sequential relays on the same band, and to disconnect the converter when operating high power on other bands. The printed-circuit boards will accommodate the input diodes if you want to include them.

Layouts for the printed-circuit boards are shown in figs. 3 and 4. These boards

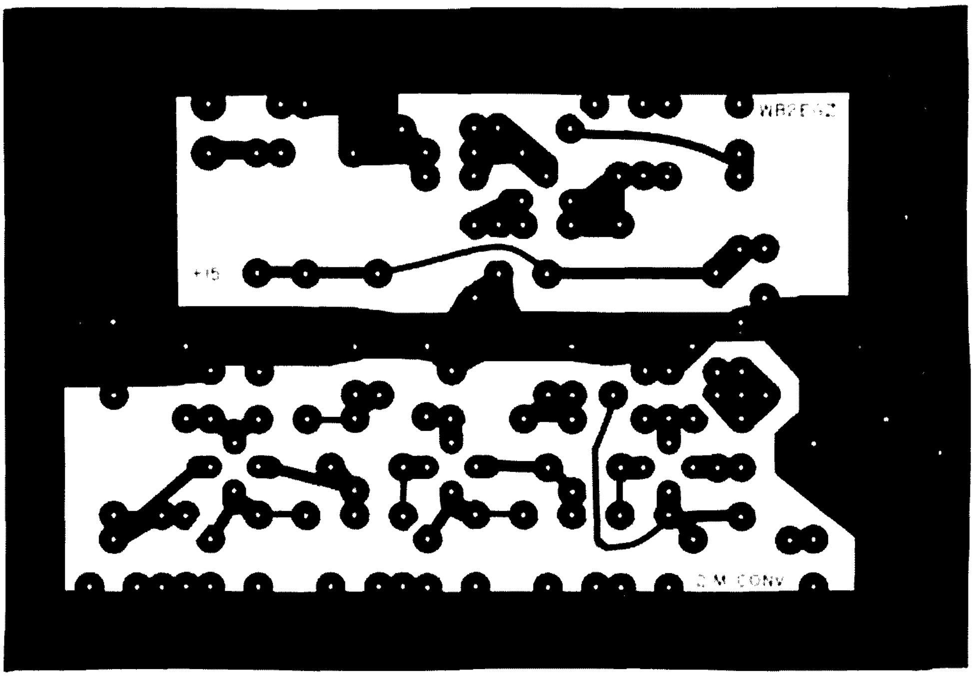
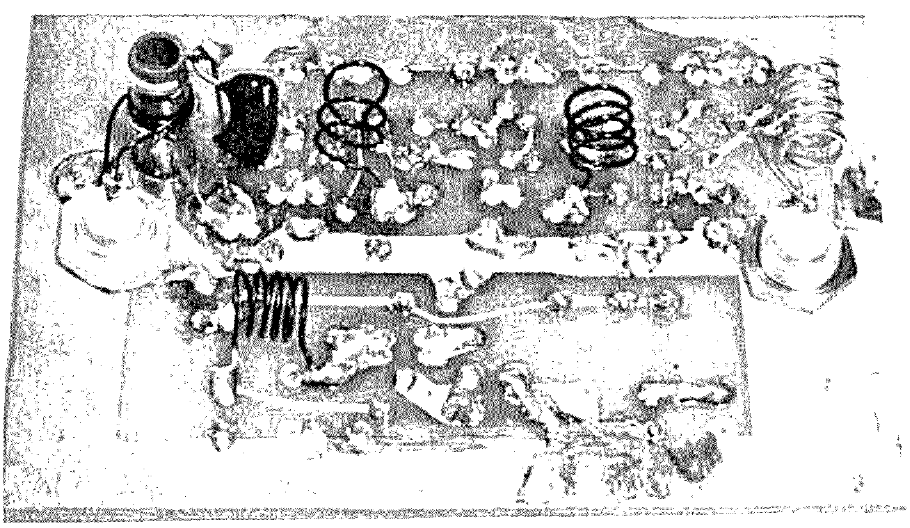


fig. 3. Full-size printed-circuit board for two-meter converter.

Bottom view of the two-meter converter. Stability of the first stage can be improved by mounting the antenna coil vertically, 90° to L2.



have been designed to include a 15-volt zener regulator and voltage-dropping resistor in case you don't have a convenient power supply. The value of the voltage-dropping resistor can be calculated on the basis of total current drain of 40 to 50 mA.

The printed-circuit design allows for a variety of components. For example, you can use an HC-6, HC-18 or HC-35 crystal holder. The trimmer capacitors may be replaced by feedthrough types if you want. And, except for L2 in the six-meter converter, the printed-circuit-type coils can be replaced by feedthrough equivalents. If you want to operate from a 12-volt source, replace the coupling resistors in the rf stages with 1.5 μH chokes in the two-meter unit, and a 6.8 μH choke in the six-meter converter.

tuneup

All tuned circuits except the output tank may be checked with a grid-dip oscillator, Dip the oscillator tank circuit a few MHz higher than the crystal frequency. Apply power and tune up on the band. In some converters the first two tank circuits must be stagger-tuned to avoid oscillation. Some broadbanding is normally desired anyway, so no shielding was included in early models. Later experiments showed that a simple shield between L1 and L2 helped stabilize the amplifier. The circuit may also be stabilized by broadbanding with a 2200-ohm resistor across L2.

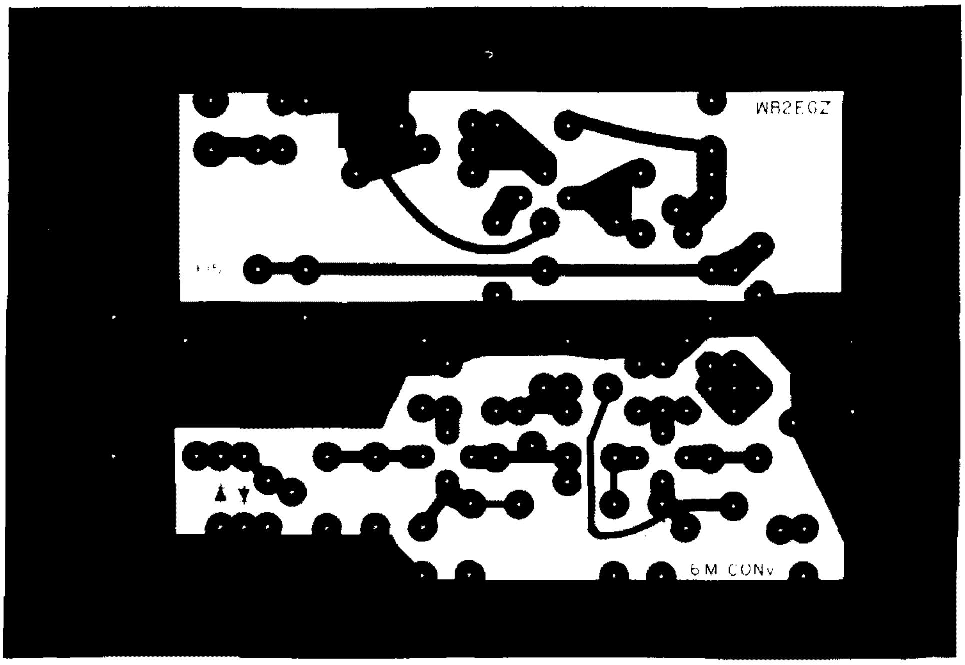


fig. 4 Full-size printed-circuit board for six-meter converter.

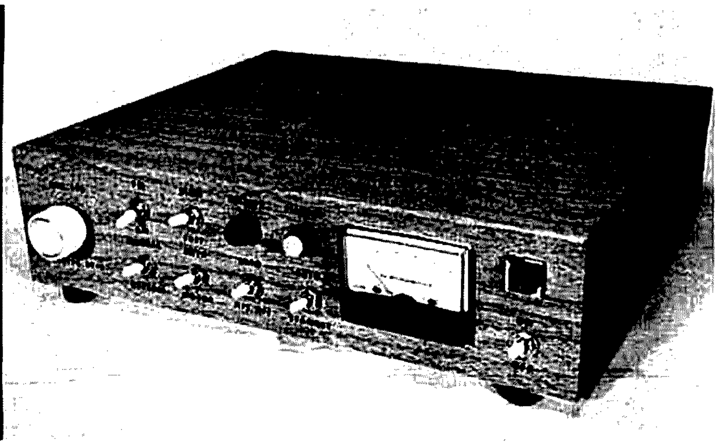
The noise figure of this design has been measured from 2.3 to 3.0 dB, depending on tuning and individual difference between components. If you select transistors for noise figure, you can obtain somewhat better noise performance. I observed noise figures less than 2 dB at an East Coast VHF Conference in 1969.

I would like to thank Mike Ward, WB2YJK, Joe Bennet, WB2FDL, and Steve Wojcik for their contributions to this article.

references

1. Donald W. Nelson, WB2EGZ, "The WB2EGZ Converter," *ham radio,* June, 1968, p. 22.
2. Donald W. Nelson, WB2EGZ, "The Two-Meter Winner," *ham radio,* August, 1968, p. 22.

ham radio



troubleshooting the ST-6 RTTY demodulator

Complete troubleshooting instructions are given along with voltage measurements and theory of operation

Irvin M. Hoff, W6FFC, Los Altos Hills, California 94022

The article on the ST-6 in the January issue[1] was somewhat general in scope. There are several topics remaining which are of interest to the serious enthusiast, including voltage charts, trouble-shooting comments and more detailed analysis of circuit operation.

voltage measurements. These were all made with a Heathkit vtvm with the standard 10 megohm dc probe. The power supply had ±12.0 volts output. On each op amp, the voltages at pin 4 were about −11.9 volts, and at pin 7, about +11.9 volts. If you are reasonably close to these measurements and the ones that follow, you should expect normal operation in that part of the circuit. Other voltage measurements are shown in table 1.

Some latitude on these measurements is acceptable. The voltage with mark signal at test point 2 might vary from unit to unit, but should be 6 to 9 volts. All measurements on U1 should be pretty close to those given. On U2 most of the measurements will be pretty close; if the voltage at pin 6 is not between 7.5 to 9

volts, adjust resistor R'D' as explained later. On U3 the measurements will be very close again, with voltages at pins 2, 3 and 6 slightly less than at pin 6 of U2. U4 should be quite close to these. On U5 the voltages at pins 2 and 3 may vary a little, but the rest should be very close. On U6 and U7 the voltages should be pretty close to those shown in table 1.

On Q1, the voltages at the base should match very closely, although those on the collector can vary some. That is, with a mark signal, you might get from 0.5 to 2.0 volts; the voltage on space will depend on your loop supply and transformer, but probably will be between 140 and 200 volts. Q2 and Q2 voltages should match closely. The voltage on the base and emitter of Q3 may vary somewhat. The voltage on the collector of Q4 may be anything from almost zero to 0.2 volts. The voltages on Q5 and Q6 should be about the same as those in table 1. On Q7 the collector voltage with mark signal will be from almost zero to 0.2 volts, and the voltage on space might be anything from 8 to 10 volts.

troubleshooting

The first thing to check is power supply output; make sure you are getting close to ±12 volts. Then start with the U1 stage. Disconnect the input to the ST-6 and balance the pot for 0 Vdc output; then connect the input and insert a mark signal. Check the voltage at test point 2; it should be around 7 to 9 volts. Check the voltage at test point 3, it should also be around 7 to 9 volts. If more than 9 or less than 7, you might want to change resistor R'D' on U2 until you get close to but not more than 9 volts. Then pick an appropriate capacitor for C'C' from table 2.

By keeping the voltage close to 9 Vdc maximum dynamic range will be obtained for limiterless linear operation. If the voltage at test point 3 is allowed to exceed 9 volts, the input to U4 could (with RTTY being received) go higher than 4.5 volts and possibly damage U4, as the maximum differential input voltage for a 709C op amp is 5.0 volts.

If the output at test point 3 is around 8 or 9 volts, then the output at pin 6 of U4 should be approximately +10.8 volts. This should cause the printer to stay in mark. With a signal, the motor should have come on within 4 or 5 seconds or less; if not, the autostart system is not working properly or has not been adjusted correctly.

theory of operation

automatic printer control. Let's approach this from two different aspects. Let's say we have been printing an incoming signal; just this instant it quit and we are getting random noise from the receiver – the motor is still running and the receive lamp is still on.

With a random noise input, the voltage at test point 2 will vary from one moment to the next, but in any event it should be less than the 7.5 volts or so you had when the signal was present. Thus, the fixed bias on the non-inverting input of U5 will be greater than that coming to pin 2 of U5. (The two 68k resistors, R66 and R67 reduce the voltage at test point 2 to one-half so that the 5.0 volt input limit of U5 will not be exceeded.) Thus U5 is now controlled by the voltage at pin 3, and the output is approximately +11 volts. This is passed by CR27 and charges the 350 μF capacitor C21 through R61.

Since C21 is in parallel with a resistor network, the capacitor takes 0.8 to 1.0 second to charge. As it charges, the voltage at terminal 3 of U6 rises to about 4.7 volts. As it goes above the 2.2 volt fixed bias level on pin 2, it takes charge of U6, which then flips to positive output of about +11 volts. This is blocked by CR25 but passed by CR24, and applied to the standby line, which causes Q1 to conduct and puts the printer into mark-hold.

At the same time, the 20-μF capacitor

table 1. Dc voltage measurements in the ST-6 RTTY demodulator.

	Op Amp U1 mark signal		Op Amp U2 mark signal		Op Amp U3 mark	Op Amp U3 space
1.	7.8	1.	8.2	1.	8.2	
2.	0	2.	0	2.	8.5	-8.5
3.	0	3.	0	3.	8.5	-8.5
4.	-11.8	4.	-11.9	4.	-11.9	
5.	-11.4	5.	-11.4	5.	-11.5	
6.	7.7 Vac	6.	8.4	6.	8.5	-8.5
7.	11.8	7.	11.9	7.	11.9	
8.	9.4	8.	8.7	8.	8.5	

	Op Amp U4 mark signal		Op Amp U5 signal	Op Amp U5 no signal		Op Amp U6 signal	Op Amp U6 no signal
2.	0	1.	7.4	8.2	1.	7.5	8.1
3.	2.0	2.	3.9	0	2.	2.2	2.2
6.	11.0	3.	3.4	3.4	3.	0	4.7
		4.	-11.8	-11.8	4.	-11.9	-11.9
		5.	-11.0	-11.9	5.	-10.8	-11.9
		6.	-10.8	11.4	6.	-10.8	10.8
		7.	11.8	11.8	7.	11.9	11.9
		8.	11.5	7.8	8.	11.4	7.7

	Op Amp U7 mark	Op Amp U7 space
1.	7.5	8.1
2.	2.3	2.3
3.	0	4.2
4.	-11.9	-11.9
5.	-11.9	-11.9
6.	-10.8	10.8
7.	11.9	11.9
8.	11.5	11.5

	Q1 mark	Q1 space		Q2 S3 on	Q2 S3 off
B	0.6	-0.7	B	11.3	12.0
E	0	0	E	12.0	12.0
C	1.0	170	C	11.9	0

	Q3 signal	Q3 no signal
B	-10.3	0
E	-9.5	0
C	-12.0	-12.0

	Q4 signal	Q4 no signal		Q5 signal	Q5 no signal		Q6 signal	Q6 no signal		Q7 mark	Q7 space
B	-0.7	0	B	-11.3	-12.0	B	-11.9	-11.3	B	0.7	-0.7
E	0	0	E	-12.0	-12.0	E	-12.0	-12.0	E	0	0
C	-0.03	-12.0	C	-11.8	+10.2	C	+12.0	-11.8	C	0.05	9.3

Test point 2: Mark signal, 170 shift discriminator 7.8 Vdc; input to ST-6 disconnected, 0 volts.

C20 no longer has any voltage to keep it charged, so its 9- to 10-volt charge starts to bleed off slowly. When it has dropped to less than −0.6 or −0.7 volts, Q3, Q4 and Q5 stop conducting, causing the relay to open and the motor to turn off.

At this time Q6 conducts, pulling about the same current through the 500-ohm resistor R47 that the relay had been pulling through Q5. Thus the drain on the power supply remains rather constant, and better regulation is possible. In fact, at the time this particular part of the circuit was developed I had not intended

table 2. Correct values of C'C' for different values of R'D' (see fig. 5, reference 1). Resistor R'D' is chosen for 9 volts at test point 3 with a mark input (see text for discussion).

R'D'	C'C'
300k	.018 μF
270k	.02 μF
240k	.022 μF
220k	.025 μF
200k	.027 μF
180k	.03 μF
160k	.033 μF

to use any type of regulation in the power supply at all.

For the next configuration we have the motor off, and there is no signal. Now we suddenly get a signal into the ST-6. The voltage at test point 2 goes to approximately 7.5 volts. This becomes around 3.8 volts at pin 2 of U5 and as this is somewhat more than the fixed bias (3.2 volts) at pin 3, the output switches from +11 volts to −11 volts. (Pin 2 is called the "inverting input," and a positive signal becomes negative at the output.) This voltage is blocked by CR27, so the 350-μF capacitor C21 starts to bleed off through R59 and R60.

As the voltage at pin 3 falls lower than about 2.2 volts, the fixed bias on pin 2 takes over, and the output of U6 flips from positive to negative output. This is blocked by CR24, so now the standby system is removed and the printer is free to follow the incoming signal. At the same time the negative output of U6 is passed by CR25, so the 20-μF capacitor C20 is quickly charged through the current-limiting resistor R55.

As this happens, Q3, Q4 and Q5 all conduct, closing the relay and turning the motor on. This causes Q6 to stop conducting, so again the current in this part of the circuit is similar to what it had been when the relay was not being used.

At this point it is worthwhile to mention how the circuit could be changed to a "fail-safe" type. The relay could be placed in the collector of Q6, and resistor R47 would then be in the collector circuit of Q5. The relay would be activated any time the motor was supposed to be off (the back contacts on the relay would be used for the motor), and if the relay or any part of the ST-6 failed, the motor would come on automatically. However, with solid-state circuits there is not too much reason to worry about a fail-safe system.

fast-slow switch. This consists of a 150-μF capacitor C22 in series with the 350-μF C21. The total capacitance in "fast" then becomes about 85 μF, and the circuit responds in about one-fourth the normal time, making attended *fast-break* operation possible, yet retaining the features of automatic printer control. Switch section S4B operates at the same time to keep the motor running. This type of circuit is not suitable for unattended operation due to the short time constants.

anti-space. This circuit is quite interesting, and some knowledge of how it works is beneficial. Basically it puts the printer back to markhold if the input signal goes to space for longer than a normal RTTY character. This also prevents the autostart from activating if a signal appears in the space channel. Therefore, if a station is playing around with his shift and going between mark and space, your machine will not run open.

With a mark signal a positive voltage will appear at the output of U4, causing Q7 to conduct. Its collector then goes to less than 0.2 volts, effectively short-circuiting the 10-μF capacitor C19. The 330-ohm resistor R40 is there to limit current through the transistor as it suddenly shorts this capacitor. The voltage at pin 3 of U7 now becomes approximately 0.1 volt or less, and the 2.5 volts fixed bias on pin 2 takes over, putting negative voltage out of U7. This is blocked by CR21 and CR22, so nothing at all happens.

With a space signal the output of U4 negative, so Q7 stops conducting and is biased off, the voltage held to −0.7 by the protective diode CR20 in its base. The 10-

μF capacitor is now charged to about 9 volts. This becomes about 4.0 at pin 3 of U7 as the network is provided to keep this less than the 5.0 volts maximum allowable input. It will take about a quarter-second for the voltage at pin 3 to build up more than the 2.5 volts fixed bias on pin 3. When it does, U7 flips to positive output, and the voltage is passed by both CR21 and CR22. CR21 goes to the standby line, and puts the printer into markhold. At the same time the voltage through CR22 goes to the automatic printer control line and starts to charge up the 350-μF capacitor C21. This will take 0.8 to 1 second, putting the system into "autostart off."

As this happens much faster than the incoming signal could discharge that capacitor, the motor would never turn on should an incoming space signal be received, and if a signal suddenly goes to space after printing authentic RTTY, the autostart would soon be disabled, and 20 to 30 seconds later the motor would turn off. The anti-space works equally well whether the limiter or autostart are on or off, or whether the incoming signal is being "straddle-tuned" due to incorrect shift.

No provisions were provided to disable the anti-space. However, if you need this feature for any reason, just short the collector of Q7 to ground. A switch could be permanently added for this purpose, if needed.

fsk output. With nothing connected to the J3 keyer jack, the voltage there should swing from approximately -35 volts on mark to +35 volts on space. Without the 12k resistor R31, this voltage would be around 70 volts or so, which is enough to destroy most germanium diodes used in typical fsk circuits. When the fsk system is plugged into J3, the current on conduction is held to 4 or 5 mA by the 8.2k resistor R30, offering suitable saturation for good keying of the transmitter.

the threshold corrector. I have not said much about this circuit, and indeed it is complex enough to merit a complete article of its own. Perhaps a few words will help you understand its operation. If a steady mark signal is present, the voltage at test point 3 will be about 8.5 volts. Let's call this 8.0 volts for the time being. This voltage would charge C6 and also be passed by CR10, eventually going through R18, R19 and CR13 to ground. Thus, the voltage at the input of the normal-reverse switch would be roughly 4.0 volts. (There is another network in the normal-reverse switch which further drops this 4.0 volts to 2.0 volts at pin 3 of U4; this will be explained in a moment.)

With steady mark of 8.0 volts, the output of the atc section is 4.0 volts and the input to U4 is 2.0 volts. If you now flipped suddenly to space, you should get the same 8.0 volts at test point 3; however it would now be -8.0 volts. This would be passed by CR11, and on to ground through R19, R18 and CR12, again putting -4.0 volts at the output of the atc system. This negative -8.0 volts would also charge C5.

The interesting thing that happens on a quick reversal such as this, however, is that the previously charged capacitor C6 now discharges, adding another -8.0 volts to the system, which is reduced to 4.0 volts by the action of R18 and R19; thus you get the -4.0 volts from test point 3, plus this capacitor discharge voltage of an additional -4.0, making a total voltage at the output of the atcc system of -8.0 volts.

With steady input of mark or space, you would have half that voltage appearing at the output of the atc circuit, but with steady reversals (RTTY characters), the voltage on the output would be approximately equal to the original input. Therefore, the new voltage at the input of U4 would be about 4.0 volts. The reason for the 220k resistor network now becomes more apparent, if you remember that the input of U4 cannot safely exceed 5.0 volts.

If copying mark-only signals but with RTTY reversals, it can be shown that the voltage at test point 3 would be roughly

+8.0 volts for mark and close to zero volts for space if you assume for a moment that the system doesn't particularly respond to noise alone. In this case you would have an on-off voltage instead of the plus-minus swing previously mentioned with normal two-channel RTTY.

I could show that the voltage at the output of the atc would go from +4.0 to about −4.0 volts and the voltage at the input of U4, instead of being ±4.0 volts with normal RTTY, would be ±2.0 volts with mark-only RTTY. Since the slicer is operated in an open-loop configuration, anything more than a few microvolts plus-and-minus will cause it to operate suitably. The important thing is to keep the information properly centered plus-and-minus, not *how much* it swings one way or the other.

The atc is a most fascinating circuit, and goes a great way in counteracting for signals that drift, for shifts that are incorrect, and even to some extent, for signals that have distortion on them when received.

An important thing most RTTY enthusiasts overlook is the diversity effect offered by such a system, as it actually samples both the mark and the space signal and uses either or both to provide proper information to the slicer. A system such as that used in the ST-6 actually offers diversity reception with only one antenna, one receiver and one converter. At one time commercial stations had to go to dual-diversity reception to allow for selective fading. This required two antennas, two receivers and two converters that fed into a diversity combiner. All this duplication was beyond the ability of most amateurs (and many commercial and military installations as well) so other means of improving the signal had to be found. The atc represents one of the best of such solutions.

Another system which is much more sophisticated but built around the same concept is called dtc (decision threshold computer) and was patented by Page Engineering for use on the Dew-line defense system. As long as the limiter is used, for all practical purposes atc works as well as dtc. With the limiter off, the dtc has some advantages, particularly in slow hand-speed transmission, but it does not work well at all if mark-only handsent transmissions are being used. (Neither system works well on slow-speed space only.)

Thus, from a practical standpoint atc was selected for the ST-6 rather than dtc. With the simple linear discriminators used in the ST-6 (and most TT/L-2 units), limiterless copy has little to offer. You also lose stability of tuning indicator information, whether using the meter or an external scope, and you lose the ability to include automatic printer control. Finally, with low voltage solid-state circuits, the dtc is not at all practical in the form most amateurs have seen it.

I did develop a dtc system for 12-volt systems, but it took a total of 8 op amps, and to be compatible, you would have to add active detector circuits that require two more op amps. All this seemed too stiff a penalty to pay for the occasional limiterless operation an individual might use. If going to that extra work, then it seemed silly to retain simple linear discriminators. As a consequence, the ST-6 is presented in a form that represents excellent performance, with practical considerations rather than going for the very complex unit that could have been designed. As an example, if you were to build a really top-of-the-line unit with sharp mark and space filters (three toroids each) then you would find it difficult to use the unit at all if somebody's shift were off a little bit. As you are no doubt aware, few amateurs actually are within 50 Hz of 850 to start with, so practical considerations almost immediately rule out using first-rate filters anyway. Other examples could be used to show why, with present circumstances, you might not want such a unit even if it was offered.

no automatic printer control. While I think you ought to build the ST-6 pretty much as presented, some people still insist they do not need the autostart

features. They would probably not want the anti-space either in this case, in order to save additional construction costs. The ST-6 was laid out so that all you need for similar performance is a loop supply, a ±12 volt supply, the U1, U2, U3, U4 and Q1 circuitry, and a switch from the collector of Q1 to ground for standby. That would do it. If that is still more than you need for RTTY work, forget the ST-6 and build the ST-5 presented in the September, 1970 issue of *ham radio.*[2]

summary

For those of you interested in learning more about solid-state circuits, this brief review should help. In any event you will better understand how the ST-6 works and what the various systems are supposed to do. It has taken me about six years of development to reach this phase, and there were many interesting thoughts picked up along the way. I feel the ST-6 will not be obsoleted or antiquated in any way for some years to come, although some designers may wish to use more exotic filters, or newer op amps, or add digital logic for autoprint.

Incidentally, Sel-cal devices[3] attach to the ST-6 with only a 10k resistor and a silicon diode. Since only a few dozen of the Selcal units have been built, write to me if you need that information. Large, easy-to-read schematics are still available from me for $1, postpaid anywhere in the USA or Canada; add $1 for air mail to other areas. Over 400 amateurs have requested schematics as of this writing, so it seems evident that the ST-6 promises to be one of the most popular RTTY demodulators of all time.

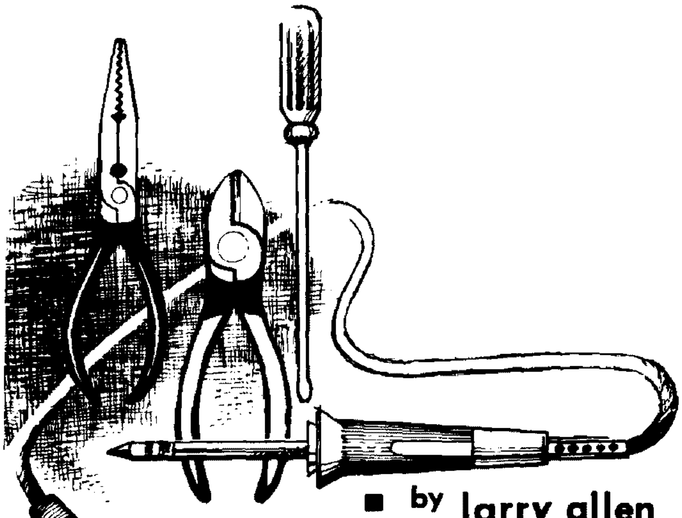
references

1. Irvin M. Hoff, W6FFC, "The Mainline ST-6 RTTY Demodulator," *ham radio,* January, 1971, page 6.

2. Irvin M. Hoff, W6FFC, "The Mainline ST-5 RTTY Demodulator," *ham radio,* September, 1970, page 14.

3. W. Mallock WA8PCK, and T. Lamb, K8ERV, "The Selcal – An RTTY Character Recognizer," *73,* May, 1968, p. 58.

ham radio



the repair bench

■ by larry allen

thinking your way through repairs

The editor sends me a lot of letters from fans of the *repair bench.* Many of them ask for help with this or that problem. Naturally, I can't answer them all. Yet, there should be some way to help you, and perhaps what I'm writing about this month will do it.

Very few hams are professional repair technicians. If they were, they'd probably get enough of electronics all day long and wouldn't fool with it as a hobby. But when something goes wrong with your gear, even knowing exactly how it's built and how it works doesn't always lead you to the defect very quickly. At those times you probably wish you were a technician trained in hunting down electronic faults.

I can't train you in a few magazine pages to be a technician. But I can show you a technique used by professionals in home-entertainment electronics.

There's one method that should be easy to learn. It's called *1-2-3-4 Servicing.* It was originated by Forest H. Belt in some books published by Howard W. Sams & Co. Although the books are not about ham gear, I could see immediately how easily the method applies to any kind of electronics. So here's how *1-2-3-4 Servicing* can be applied to equipment on your ham repair bench.

thinking in sections

The important thing about 1-2-3-4 Servicing is the way you think. You have to think of electronic equipment in an orderly fashion. For that purpose, all electronic devices are split up into four logical divisions. They are sections, stages, circuits, and parts.

First, you think *sections.* Take a ham receiver, for example. It has an rf section, i-f section, audio section, and power supply section. These are sketched for you in fig. 1. You might be tempted to break it down further, but these are the breakdowns you use for 1-2-3-4 Servicing.

A transmitter divides up into the speech or audio section, rf generating section, modulated rf section, rf power section, and power supply section. If the transmitter is ssb, there is also a frequency-change section. Transmitter sections are illustrated in fig. 1, too, and the comparison with the sections of a receiver may help you see what sections really are.

The distinction that identifies a section is simple: *A section handles only one kind of signal.*

In a receiver, for example, the rf section handles incoming stations. The i-f section amplifies the station signal only after it has been heterodyned down to the intermediate frequency of the receiver. The audio section handles voice signals alone, after they have been demodulated from the i-f signal.

In a transmitter, as long as audio (voice or speech) is by itself the audio section handles it. The rf signal, as long as it's alone, is in the rf-generating section. Once modulated, the signal is handled by a different section. If the transmitter is ssb, a frequency-changing section takes the modulated signal and changes its frequency to the transmitting frequency. Once the modulated signal is at the output frequency, an rf power section boosts it.

That's how you can identify sections in any kind of electronic gear. When the signal changes character, consider it as going into a different section.

stage by stage

The second division you think of for 1-2-3-4 Servicing is *stages*. Sections are divided into stages.

For example, the rf section of a ham receiver has at least three stages. They're shown in fig. 2A. There may be more than one rf amplifier stage. A stage usually comprises one tube or one transistor and the parts that go with it. A stage either generates or amplifies or alters a signal.

Inside this rf section, the oscillator stage produces a plain rf signal. The rf amp gives a boost to the signal picked up by the antenna (which can also be considered a stage). Both signals go to the mixer. That stage changes the character of the rf signals, mixing them together and coming up with a signal at some intermediate frequency (their difference).

The i-f signal still has the station modulation that was on the rf signal. But the character of the signal has changed. It takes a different kind of stage to handle it. So the mixer stage has introduced the signal from the rf section into the i-f section.

The mixer stage, though officially part of the rf section, is an *interface stage*. It interfaces the rf section to the i-f section. The stage that creates the character change in the signal is an interface stage.

The detector or demodulator following the i-f amplifier stages is an interface stage. It extracts the voice signal that is part of the i-f signal. Once demodulation takes place, there is no more i-f signal — only the voice or audio signal. So the detector stage interfaces the i-f section to the audio section. Amplifier stages in the audio section handle only audio signals.

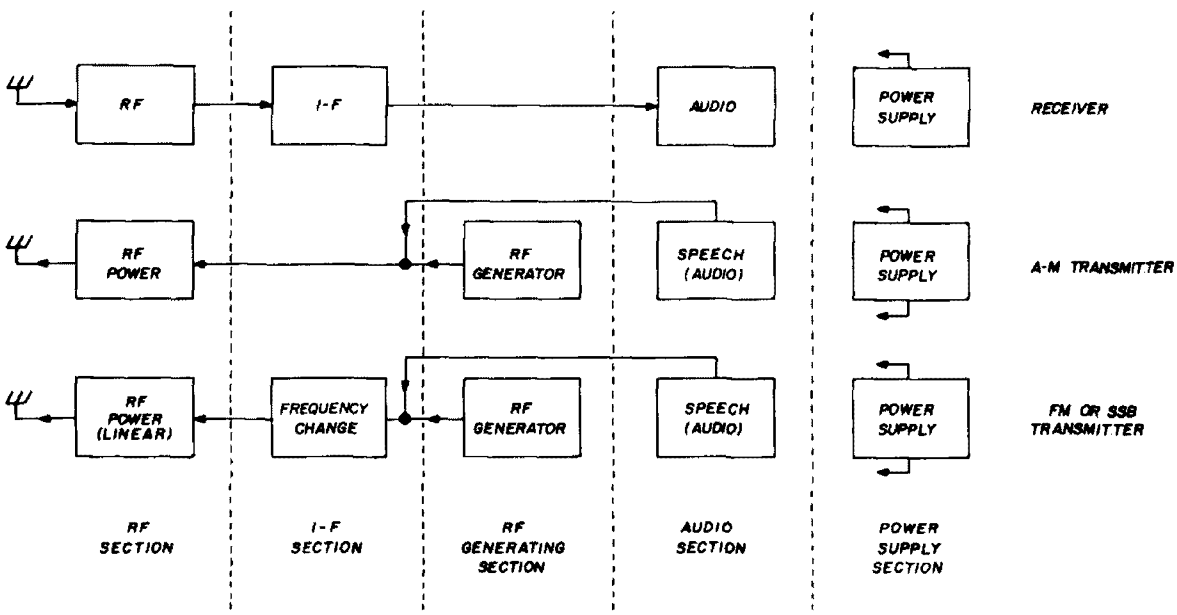


fig. 1. Sectionalizing ham equipment helps track down faults. Sections of different kinds of equipment are amazingly alike — if you learn the sections of one it's easy to divide up other electronic gear in this way.

If you want to carry the idea further, the speaker is a stage of the audio section. Actually, it's an interface stage between the audio section and the air. It converts the electronic audio signal into audible sound waves. That's what any interface stage does: change the nature of the signal between two sections.

An interface stage is always considered part of the section preceding. Sometimes it takes signals from more than one stage, as the mixer does in the rf section of a receiver.

Or, a stage may interface more than two sections. Fig. 2B has an example of this. This is the stage-by-stage division of an ssb transmitter. Stages are grouped

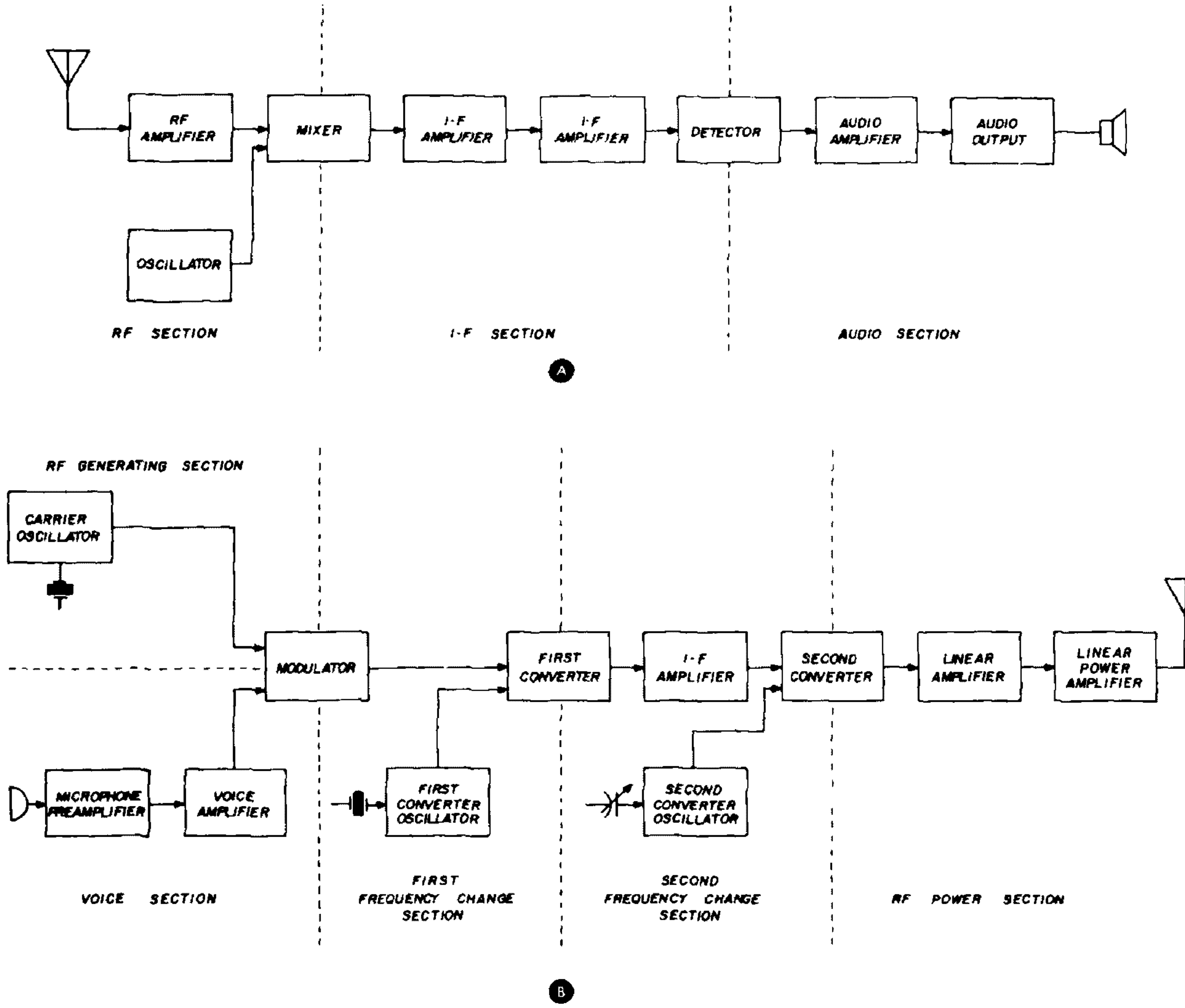


fig. 2. Sections divided into stages. For troubleshooting your work is cut down tremendously if you only have to test the stages in one section.

into sections so you can see the relationships. The voice section and rf-generating section are interfaced with the first frequency-change section by the modulator stage.

The voice section has two stages. The rf generator section (which can be called the *carrier* section in an ssb transmitter) also has two: the carrier oscillator and the modulator. It's customary to consider the modulator as an rf stage rather than a voice or audio stage, even though it involves both kinds of signals. The modulator takes the two signals and transforms them into a modulated rf signal. That calls for a different section, so the modulator is an interface stage.

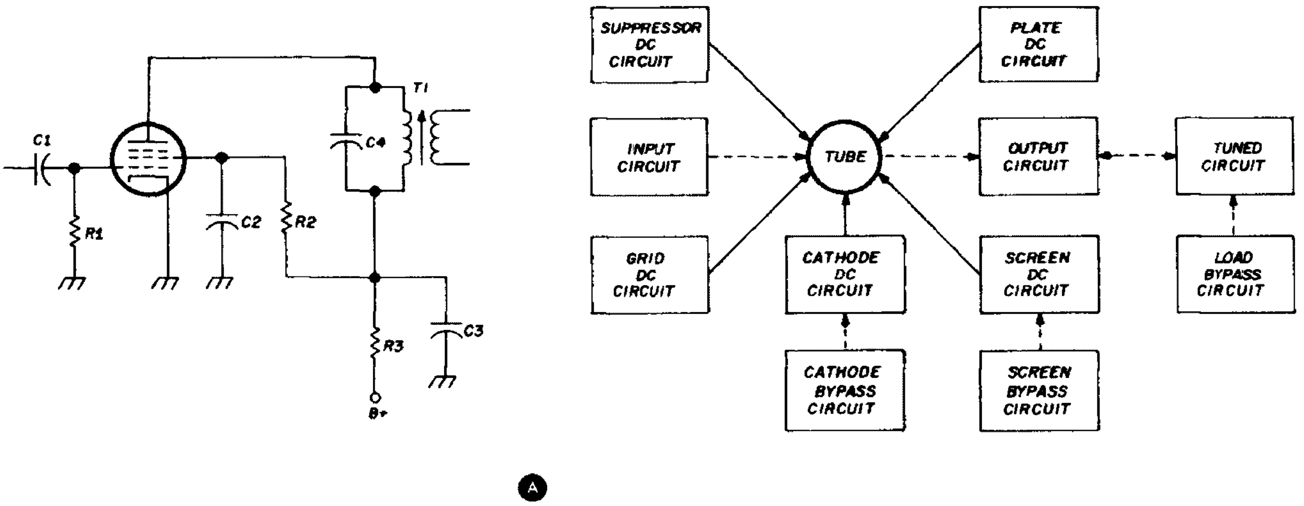
The first converter stage interfaces the first frequency-change section to the second. This ssb transmitter uses double conversion to arrive at the output frequency. The amplifier stage in the second frequency-change section is called an i-f amp. (In a few transceivers, it's the same i-f amp used during receiving.)

The modulated i-f signal goes to the second converter, the interface stage between the second frequency-change section and the rf power section. A second conversion oscillator stage furnishes the unmodulated signal for conversion. The output of the second converter is a modulated rf signal at the output frequency.

All that remains is to amplify the output signal to full output power of the transmitter and feed it to the antenna. The rf power amps and antenna are stages of the rf power section.

what a circuit really is

That should give you a good idea how



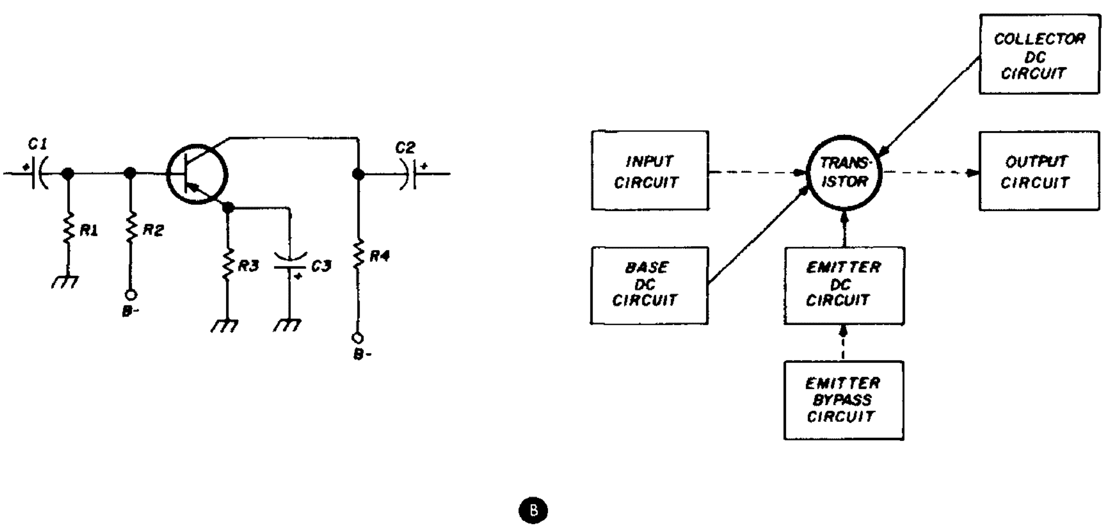


fig. 3. Identifying circuits in tube and transistor stages. Among circuit not shown here are feedback circuits (such as in an oscillator) and power-supply decoupling circuits (which aren't always drawn on schematics).

to think in sections and stages. Be sure you understand these divisions, because they're important in applying 1-2-3-4 Servicing.

The third logical division of electronic equipment is the *circuit.* Many hams – and a lot of technicians, too – think of circuits as being the things just described as stages. That's confusing. Actually, stages are made up of circuits.

Remember, a stage is a tube or transistor and the parts that go with it. Those parts are connected into *circuits.* There's an input circuit and an output circuit; a grid circuit, plate circuit, cathode circuit, and perhaps a screen circuit; a base circuit, emitter circuit, and collector circuit; there may be a feedback circuit, a tuned circuit, and so on. Any given tube or transistor stage has several circuits.

What you name the circuits depends on how you're considering the stage. For example, the tube and transistor stages in **fig. 3** have both dc and signal circuits identified for you. Solid arrows are for dc, and dashed arrows for signals.

The input circuit in both stages is C1 and R1. C1 is the input coupling capacitor, R1 the input load.

The output circuit for the tube is T1, C4, and C3. T1 and C4 are a tuned circuit that is the main load, and C3 is the load decoupling or bypass part of the circuit.

At the same time, the plate dc circuit

includes the primary winding of T1. The whole plate dc circuit is through R3 and T1. Thus some components must be involved in your thinking for more than one circuit.

Indeed, some components are in more than one circuit. Resistor R3, for instance, is in the screen dc circuit as well as in the plate dc circuit. In troubleshooting, you'd naturally expect a bad R3 to affect both dc circuits. And it would.

By the same token, R1 in the transistor stage is part of the input circuit for signals. Yet, at the same time it's part of the base dc circuit; it's part of divider R1-R2 which sets base bias. Resistor R4 is the output load, and is also in the dc collector circuit.

Two bypass or decoupling circuits are invisible in **fig. 3B**. Anytime a B-plus or B-minus connection is shown in a schematic, you must remember that it includes a filter capacitor, which is usually in the power supply. But it is part of the signal circuit of any stage fed from that leg or branch of the power supply.

When troubleshooting, don't ignore this bypass circuit. In **fig. 3B** a filter capacitor is part of the output load circuit, forming the ground return for signal at the bottom of R4. The bypass capacitor at the B-minus end of R2 makes R2 a part of the input load, in parallel with R1.

what makes up a circuit?

Of course the answer to that question is *parts.* You've already seen that, in my description of what a circuit is. And parts are the fourth division of electronic equipment.

One secret of 1-2-3-4 Servicing is this manner of thinking. You consider all electronic equipment in terms of these four divisions. You think of a receiver or transmitter as divided into sections. The sections, you think of as divided into stages. Stages are in turn divided into circuits, and those are divided into individual parts.

The four steps of thinking are, therefore: sections, stages, circuits, and parts.

Once you've learned to think of your ham gear as made up of sections, stages, circuits, and parts, you can use those divisions in an exceptionally logical way when trouble occurs. Here's how the 1-2-3-4 Servicing method works.

The first step is *diagnosis.* You diagnose which section of the faulty instrument has the trouble in it. There are three main helps to diagnosis.

One is to know the equipment. Be familiar with it. Study the schematic and read the instruction manual. Read books about it. Gain knowledge of it through experience. Once you're familiar with a receiver or transmitter, you know what sections it has. That aids diagnosis. For example, if a transmitter puts out rf, but not modulated, you know the trouble must be in the voice section.

Another approach to diagnosis is inspection. *Look* inside the unit. *Listen* to a receiver. Or listen to the relays of a transmitter. Or listen for some unnatural sound like a resistor or transformer frying or something arcing. *Sniff,* and sometimes *touch* – perhaps to sense overheating. In other words, use your senses to inspect the unit. They may help you diagnose which section has trouble.

And of course you diagnose by observing symptoms. Is the unit completely dead? Does some portion of it work? Some one function not work right? Do the operating controls work as they're supposed to? How about service adjustments? Such analysis of symptoms can point strongly to which section of a transmitter or receiver is bad.

the second step

Once you diagnose the faulty section, your next step is to *locate* the defective stage. Since you already know the faulty section, you have only a few stages to check. Instead of checking a whole setfull of stages, you check only two or three – those in the section you diagnosed.

You might locate the faulty stage by observing symptoms, or by noticing how certain controls react. But more likely you'll use instruments.

You already know there are two ways to look at stages. They are dc-operated, and they handle signals. You can test them for either kind of operation. You can inject a signal at the input of a stage and see how it travels through the remainder of the set. Or, you can use a tracer to see how far a signal gets, to see what stage stops it or distorts it or somehow fouls it up.

Or, you can measure dc voltages on stages in the section you've diagnosed as faulty. If you find a stage with dc voltages upset, you've located the faulty one. (One thing about this: if several transistor stages are dc coupled, the first one you measure may not be the truly bad one. You may have to "inject" dc voltages at strategic points to find out which stage is causing the trouble.

isolating the circuit

The third step of 1-2-3-4 Servicing is *isolating* the defective circuit. You may have done it already with some of your locating procedures.

For example, you may have traced a signal in a receiver as far as the collector of one stage but find it missing at the base of the next. You've located the faulty stage or stages; but, what's more, you've also isolated the faulty circuit. It's the circuit between the collector of one stage and the base of the next. That can only be the coupling circuit.

But even if you have no idea what circuit is bad, you have only a few to test. That's because you've eliminated those in other stages and sections by the first two steps. You now have to test only the circuits in one or at most two stages.

Signal tracing or injection both work for input, output, and coupling circuits. You can try adjusting tuned circuits; if they don't respond, they must be faulty. You can put a signal across decoupling and bypass circuits; the signal should be almost wiped out. Or, you can measure signal across them; it should be near zero.

You can use dc voltage tests. In a tube stage, you measure voltages in the plate, screen, grid, and cathode circuits. Keep in mind as you do that the grid affects all those others. If the voltages all are wrong, check for a problem in the grid circuit.

table 1. Basic steps to thinking 1-2-3-4 in servicing electronic equipment.

step	action	division
1	diagnose	sections
2	locate	stages
3	isolate	circuits
4	pinpoint	parts

In a transistor stage, the collector, base, and emitter dc circuits are the ones to check voltages in. Remember that the base controls collector voltage through its influence on current through the transistor.

If the transistor is a field-effect type, the voltages to measure are at the drain, the source, and the gate. The gate affects the drain voltage through its control on current through the channel.

pinning down trouble

The fourth and final step in 1-2-3-4 Servicing is *pinpointing* the faulty part. The job has by now been simplified almost to the point of no effort. Once you have the faulty circuit isolated, there are very few parts to think about. Even if you aren't sure which of two circuits is bad, the number of parts is small.

You can extend the tests you used in step three. You can make signal and dc tests that pinpoint the faulty part. You may have already done so during step three. For example, when you isolated the faulty coupling circuit, there's only one part so it's the bad one. That happens with other circuits, too.

But there are so few parts involved in this final step of 1-2-3-4 Servicing, you can freely succumb to individual parts tests. You've gained so much efficiency through the first three steps, this may be the quickest way to finish up. You can test most parts with your voltmeter or ohmmeter. (If you don't know how, I can tell you in a later repair bench.) Or, if you have a resistor-capacitor tester, a transistor tester, etc., use them.

Another fast way to test ordinary parts is by substitution. You can probably find what you need in your junk box.

table 2. Steps for 1-2-3-4 servicing.

DIAGNOSE (section)	A. Know the equipment	1. The schematic 2. Instruction manuals 3. Experience 4. Study books
	B. Inspect — Inside and out — Off and operating	1. Look 2. Listen 3. Smell 4. Feel
	C. Observe symptoms	1. Dead? 2. Works in part? 3. Operating controls? 4. Service adjustments?
LOCATE (stage)	A. Observe symptoms	1. Dead or operating poorly 2. Controls and adjustments
	B. Signal tests	1. Injection — Signal generators — Finger (tubes only) — From similar set
	C. Voltage tests	2. Tracing — Oscilloscope — Vtvm and probe — Signal tracer
ISOLATE (circuit)	A. Signal tests	1. Input circuit 2. Output circuit 3. Bypass circuits 4. Tuned circuits 5. Feedback circuits
	B. Dc voltage tests	1. Tube — Plate — Screen — Cathode (Grid affects these) — Grid 2. Bipolar transistor — Collector — Base — Emitter 3. Field-effect transistor — Drain — Gate — Source
PINPOINT (part)	A. Signal tests (parts in signal circuits)	1. Tracing 2. Injection
	B. Dc voltage tests	1. In-line voltage tests (Use Ohm's and Kirchhoff's laws) 2. Tests in nearby circuits (For parts connected between) 3. Remember interaction — Grid affects plate, screen, cathode — Base affects collector and emitter — Gate affects drain and source
	C. Individual parts tests	1. With special testers 2. With volts and ohms tests
	D. Substitution	(Limit to common, inexpensive parts)

Or, some manufacturers make substitution testers (fig. 4) that include capacitors and resistors of many values, some diodes, some electrolytics, and so on. Except for expensive parts, substitution is a good bet, even if you have to keep a small stock of typical values.

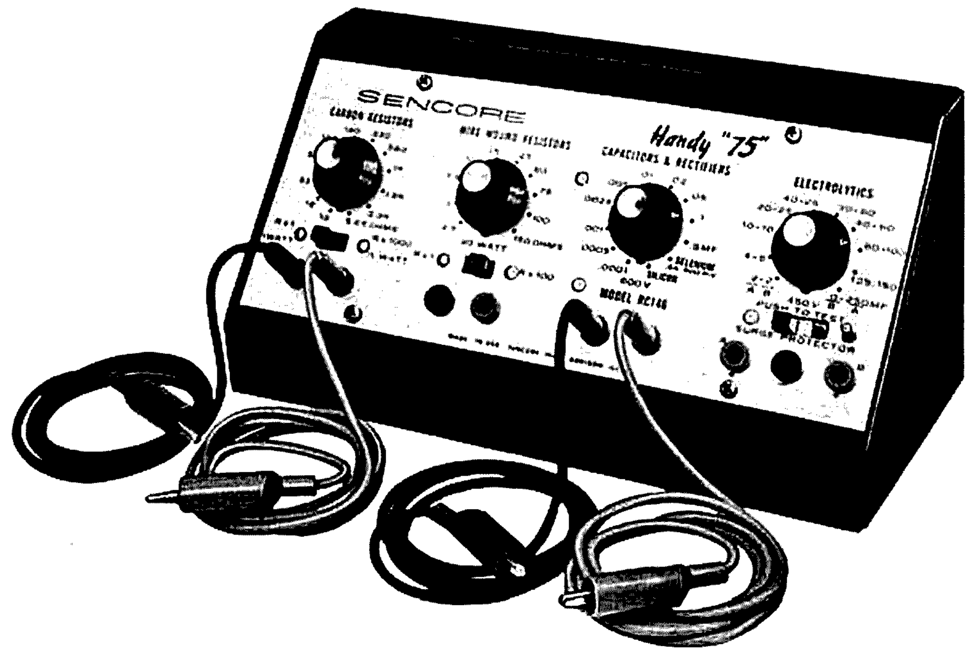


fig. 4. Commercially built parts substitutor saves carrying an inventory of small parts for test purposes.

summing up

Now you can understand the chart in **table 1.** If you looked at it before reading how to apply 1-2-3-4 Servicing, it may not have made much sense. But it summarizes the steps of 1-2-3-4 Servicing as you apply them to troubles in your own equipment.

There's nothing mysterious about this method. If you're really efficient at finding trouble that occurs in your gear, you probably already use some version of this technique. If not, however, this logical approach can make troubleshooting easier than you ever thought possible.

To jog your memory for the method and how to apply it, you may want to cut out or copy the more complete chart in **table 2.** It gives you a thorough outline of the whole process and the means by which you can accomplish it. With it, you can apply 1-2-3-4 Servicing to any piece of ham gear you own.

next time . . .

Speaking of letters from you readers, I've really got a lot of them about using a sweep generator down in the low frequencies. In the very next *repair bench* I'll tell you how.

To get ready for it, I suggest you go back and review the earlier one about using a sweep generator and scope. I won't repeat much of the general information that tells how to set up the generator and scope for sweep alignment. That way, I spend most of my next *repair bench* explaining the techniques of sweeping at low frequencies and how to make use of it.

(Which all reminds me to remind you: If you have any other special ham repair problems you want covered in this department, it takes a letter to let me know.)

ham radio

adding incremental tuning to your transceiver

This simple varactor-controlled circuit can be added to any transceiver to provide the operating benefits of incremental tuning

Michael J. Goldstein, VE3GFN, 22 Kingswood Road, Toronto 13, Ontario

There you are, sitting on 14040 kHz with your transceiver humming. You're waiting for a call from VU2JN, along with six-dozen other guys in the pileup. He's working them 2-kHz down from his transmitting frequency, and your dial hand has muscle fatigue from cranking back and forth – move down 2, give him a quick call, back up 2, listen . . . down again and call, etc. What you need is receiver incremental tuning; no problem with the little circuit shown here.

incremental tuning

What you want is a system where you can set your main tuning dial and leave it; then, by using an independent control, be able to vary operating frequency slightly to either side of the main-dial setting, either while receiving only (RIT), while transmitting only (ITT), or while both transmitting and receiving (IRTT). With these added features you can check the band around your main frequency, and return to the original frequency with the snap of a switch. If you want, you can move your QSO off a busy channel without losing the original channel setting, or listen to two QSOs without touching the main tuning dial. In fact, you can do all the things only the more expensive transceivers allow for.

Let's look at what is involved as far as including incremental tuning in your transceiver. A typical oscillator tank circuit is shown in fig. 1. After L1 and C1

have been set for the desired frequency limits of the oscillator, capacitor C2 is used to adjust output frequency. Therefore, C2, which is connected to the main tuning dial, is used to control the receiving and transmitting frequency of the transceiver.

incremental tuning circuits

If you put a small variable capacitor (typically 10 pF) across the tank circuit as shown by the dotted lines in fig. 1 (C3), set this additional capacitor at half mesh, and adjust the oscillator for the proper frequency limits, the circuit would operate exactly as before, with one exception: after setting the main tuning dial, the operating frequency could be varied slightly to either side of the main dial setting with C3. When C3 was returned to the half-mesh position, the operating frequency would be restored to the main-dial setting.

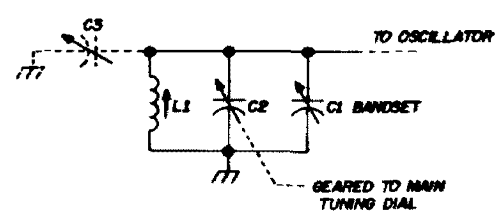


fig. 1. Simple, though impractical, incremental-tuning scheme uses added trimmer capacitor C3 across the vfo tank circuit.

However, although this is a simple way of obtaining incremental tuning, it is impractical. The same job can be done much more conveniently with a variable-capacitance diode or varactor. The varactor diode is essentially a variable capacitor that is controlled by a dc voltage instead of a rotating mechanical shaft. The capacitance range of these diodes, when operated over their rated voltage range, can be greater than 100 pF. If a 100-pF varactor was installed across the oscillator tank circuit, the result would be the same as an extra bandset capacitor, However, the effect on frequency change would be far too great.

This small problem can be solved by placing the varactor in series with a small trimmer capacitor as shown in fig. 2. The trimmer capacitor is then used to control the amount of effect diode capacitance has on oscillator frequency. If you use a 10-pF trimmer, and a 200-pF varactor, the maximum capacitance added to the tank circuit can be no greater than 10 pF.

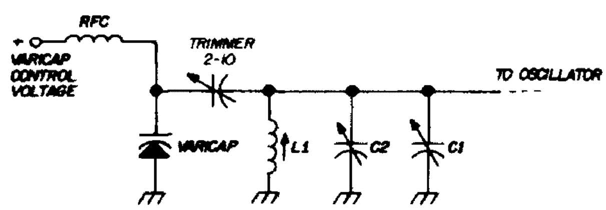


fig. 2. Varactor diode in place of C3 in fig. 1 provides practical means of obtaining incremental tuning.

transceiver modification

To incorporate incremental tuning in your own transceiver, locate the bandspread and bandset tuning capacitors. Then install the varactor diode, trimmer capacitor and rf choke as shown in fig. 2. Make sure the new components are mechanically rigid – you don't want to add instability to your transceiver's output frequency.

When soldering the diode into the circuit, protect it with heat sinks on the leads. Also, when installing the diode, make sure the anode of the varactor is grounded; the device must be reverse biased to operate correctly. The cathode end of the diode is usually marked with a white ring.

When all the components have been installed and checked, adjust the varactor bias voltage slightly more negative than the center of the recommended operating voltage range. This will lower the oscil-

lator frequency slightly, so you must adjust the bandset capacitor to re-establish main dial calibration. *Do not* touch the oscillator coil.

change of frequency with rotation; use a high-quality potentiometer to eliminate problems with noise and frequency instability.

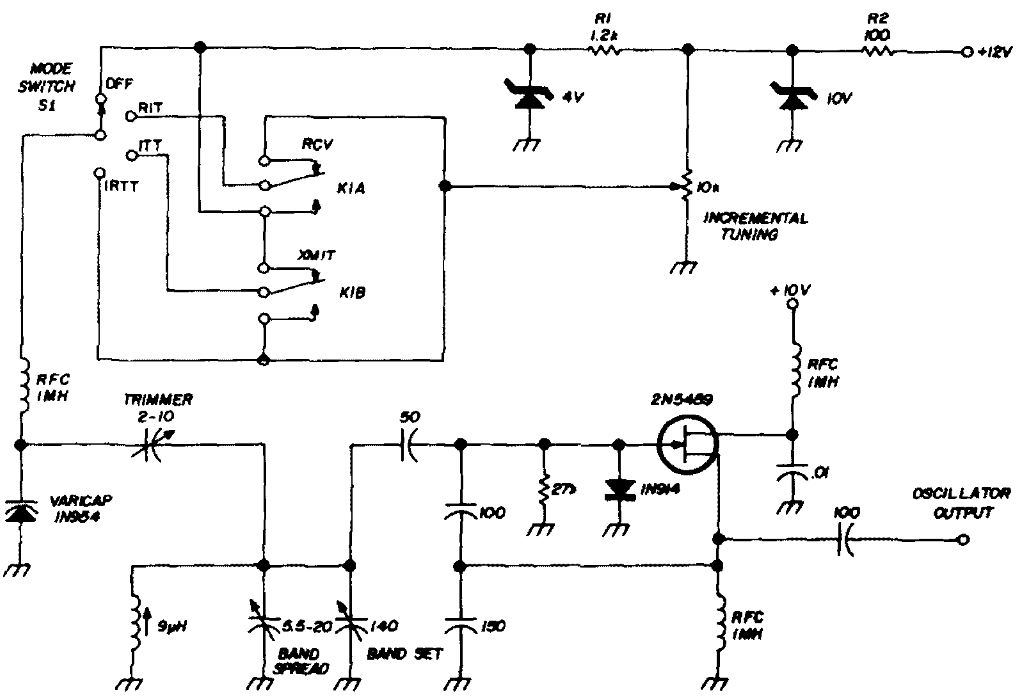


fig. 3. Complete incremental tuning circuit, as added to a 5.0 to 5.5 MHz vfo. Resistors R1 and R2 must be calculated for proper voltage drops if B+ is greater than 12 volts. Relay K1 is shown in the receive position; it must be switched with the antenna-change-over relay. With S1 in position 1, incremental tuning is off; position two, receiver incremental tuning; position three, transmitter incremental tuning; position four, receiver and transmitter incremental tuning.

The varactor's change of capacitance with voltage is not a linear function, so the "mid-point voltage" must be on the low side of the diode's voltage range mid-point, i. e., 4 volts for a zero to 10-volt varactor. Also, since the mid-point voltage of the varactor is used as a reference level for re-calibrating your main tuning dial, the bias source should be zener regulated as shown in **fig. 3.** This is especially important, because any change in transceiver operating voltages will effect the output frequency.

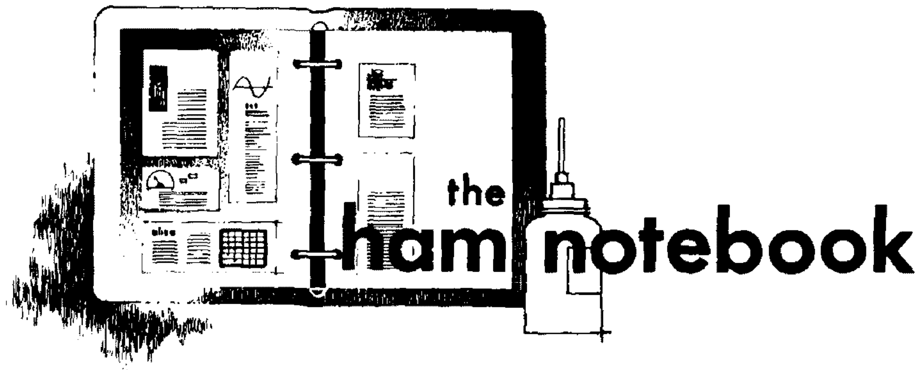
The complete incremental-tuning circuit, with all regulation and control circuits, is shown in **fig. 3.** This is the circuit for a 5.0 to 5.5 MHz solid-state oscillator that I used to test this incremental tuning scheme. The potentiometer in the circuit should be a *linear* type to provide a linear

alignment

Vary the varactor bias voltage across its full range and see how much frequency variation you get. A total of 10 kHz is ideal. If you get too much variation, decrease the series trimmer capacitance; if you don't have enough variation, increase trimmer capacitance. Each time you change the trimmer setting you must reset main-dial calibration with the band-set capacitor.

By alternately adjusting the trimmer and bandset capacitors, you can obtain the amount of incremental tuning you want, while retaining main-dial calibration. A digital frequency counter makes this adjustment very easy, but a BC-221 or external calibrated vfo will work as well.

ham radio



measurement of electrolytic capacitors

Recently I had occasion to measure the value of an electrolytic capacitor. Since test equipment was not available, I devised a simple method using a resistor, a voltmeter and a stop watch. The capacitor was connected as shown in fig. 1.

The switch should remain closed for several minutes before making the measurement. This permits the electrolytic to form and stabilize. This is particularly important if the capacitor has been out of use for a long time. Resistor R1 is included in the circuit to limit the initial surge in charging current; its value can be on the order of 1000 to 5000 ohms.

When the switch is closed, read the voltmeter. Make a note of this reading (V_B). Now, mark the voltmeter at half this reading. This can be done by simply laying a strip of paper over the voltmeter scale at the $0.5V_B$ point. With a stop watch or watch with a sweep second hand, starting with the instant the switch is opened, measure the time required for the voltage to drop to the $0.5V_B$ point. Resistor R2 should be adjusted to produce a discharge time in excess of 30 seconds; this increases accuracy since it permits easier timing.

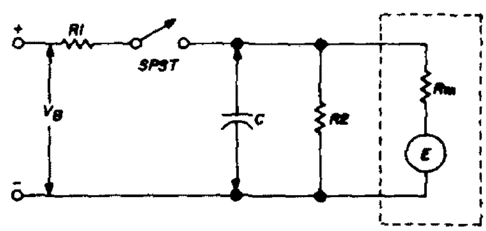


fig. 1. Circuit for determining capacitance of electrolytic capacitors. After measuring discharge time as discussed in text, capacitance can be found from nomograph in fig. 2.

The value of the capacitor can be calculated from the following equation

$$C = 1.44\,(t/R) \qquad (1)$$

where t is in seconds, R in megohms, and C in microfarads (see fig. 2). The value of R in eq. 1 is the parallel equivalent of R2 and R_M, the internal resistance of the meter (eq. 2).

$$R = \frac{R2\,R_M}{R2 + R_M} \qquad (2)$$

The resistance of a voltmeter is usually given in terms of "ohms per volt." If, for example, your voltmeter is rated at 20,000 ohms per volt, when set on the 100-volt scale R_M is 2 megohms (100

volts x 20,000 ohms = 2 megohms). Since R2 may be on the order of 1 megohm, voltmeter resistance cannot be neglected if you want accurate results.

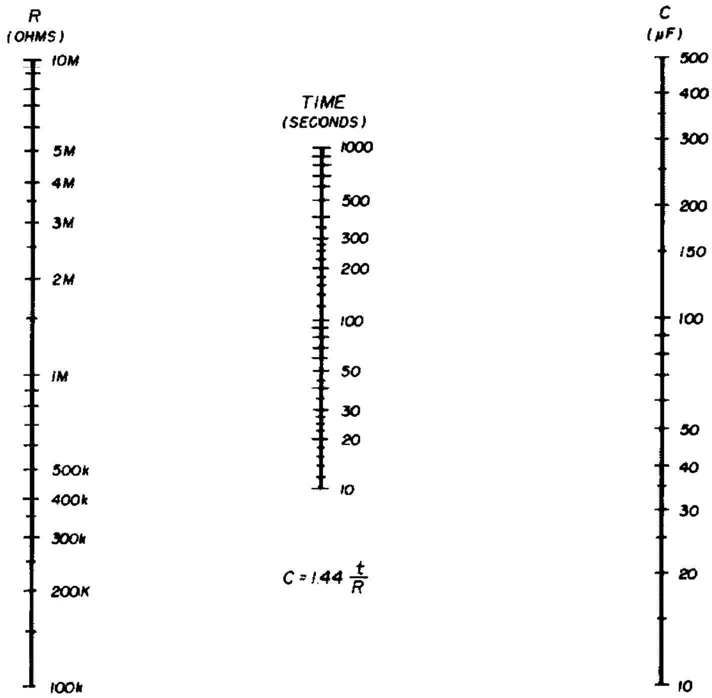


fig. 2. Nomograph for determining capacitance of electrolytic capacitors using method discussed in text.

Although this method of measuring capacitance can be used with any type of capacitor, on small-value units it's difficult to measure discharge time accurately enough to obtain meaningful results.

Edwin L. Clark, W2NA

blower maintenance

Many items of electronic equipment have blowers that move air across heat-generating components. The squirrel-cage blower seems to be popular in transmitters and transceivers, probably because these blowers are relatively quiet. Over a period of time, however, dirt builds up on the impeller in squirrel-cage blowers. Blower efficiency can gradually deteriorate without your being aware of it if dirt is allowed to accumulate. The result can be disastrous, especially if the blower is used to cool high-power transmitting tubes.

One remedy is to inspect the blower periodically and remove the impeller to give it a good cleaning. This usually involves considerable work, especially if the blower is located in an inaccessible place. Here's an effective preventive maintenance measure that not only keeps dirt buildup to a minimum, but eliminates the chore of removing the blower impeller for cleaning.

The intake port for the blower on my transceiver, a National NCX1000, is on the rear of the cabinet and flush with the vertical surface of the rear wall. After several hours of operation, I noticed that the blower impeller vanes were loaded with dirt. I used an old toothbrush to remove most of the grime, but it was necessary to remove the impeller from the blower to do a good cleaning job.

The answer to the problem consisted of a filter made from a piece of polyurethane foam about ¼-inch thick, which I cut to cover the blower intake port. Poly foam is available at retail sources that supply material for do-it-yourself furniture makers. It's used for filling seat cushions, and is available in several thicknesses. I secured the filter over the blower intake port with masking tape so it could be removed easily for cleaning.

Some types of poly foam have a more dense structure than others, so it's a good idea to test the filter before taping it into place. If the blower speed decreases appreciably, the foam is either too dense or too thick. A little experimentation will produce the right combination for effective filtering consistent with maximum blower efficiency.

I checked my blower and filter after about 90 hours of operation. The outside of the filter was coated with a rather thick deposit of dirt, but the impeller and blower housing were spotless. It was a simple job to remove the foam filter and wash it in warm water and detergent. The entire operation took about ten minutes, and the filter can be used indefinitely.

Alf Wilson, W6NIF

binding 1970 issues of ham radio

Many readers make their copies of *ham radio* into a bound volume each year. They will perhaps be surprised to note that the January and February 1970 issues are a little larger (8½ x 9½ inches) than from March 1970 on (6-1/8 x 9-1/8 inches).

Fortunately, there is no real problem. The actual printed portion of each page was left unchanged when the page size was reduced. The borders were merely narrowed. Thus, it is an easy matter to have a local printer retrim the two early issues to the new size and proceed with the binding.

The publishers of *ham radio* will be doing the same with the 1970 bound volumes which they produce.

Skip Tenney, W1NLB

improved power supply

Miniaturized power supplies for small-signal solid-state devices often present special problems, including heat reduction and adequate filtering within a limited space.

If an indicator light is necessary, an incandescent lamp across a secondary winding may be detrimental to the available rectified power output since the filament often dissipates a sizeable percentage of a small transformer's output. A 1/25-watt NE-2 neon indicator across the transformer primary will reduce heat as well as providing additional power for the load.

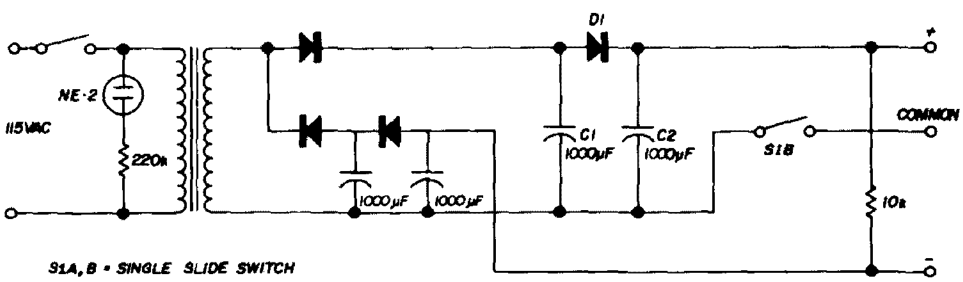


fig. 3. Improved power-supply circuit.

When a filter reactor or resistor is used in a conventional filter, both negative and positive excursions of the ripple voltage at the first capacitor are transferred to a degree to the second filter capacitor, and hence, to the load. Filtering may be improved by replacing the filter reactor or resistor with a garden variety power diode, D1, as shown in fig. 3. In this application D1 performs a gating function, allowing only the positive ripple components to reach C2, resulting in smoother filtering. If a heat-producing series resistor is replaced with the diode, cooler supply operation will result, although ripple will increase slightly.

Also shown in fig. 3 is a simple method for supplying both + and − voltages from one transformer. This arrangment is useful for powering operational amplifiers, cascode circuits and digital devices requiring both polarities.

Gene Brizendine, W4ATE

grid dipping transmission lines

Most amateurs are familiar with the many uses of the grid-dip oscillator. If a gdo is coupled to a tuned circuit made of lumped constants (fig. 4A), the meter will show a dip at resonance. Similarly, a gdo coupled to a circuit composed of distributed constants, such as a half-wave antenna (fig. 4B) or a half-wave section of transmission line (fig. 4C), the meter will dip at the resonant frequency of the antenna or transmission line.

The lowest frequency at which the gdo gives an indication of resonance is at a

half-wavelength in the case of a Hertz or dipole antenna. When grid-dipping transmission lines, a propagation-velocity factor must be used. This factor is 0.82 for typical TV twinlead and 0.66 for commonly used coax cable (figs. 4C and 4D).

When grid-dipping antennas or sections of transmission lines, either of two methods may be used. The usual method

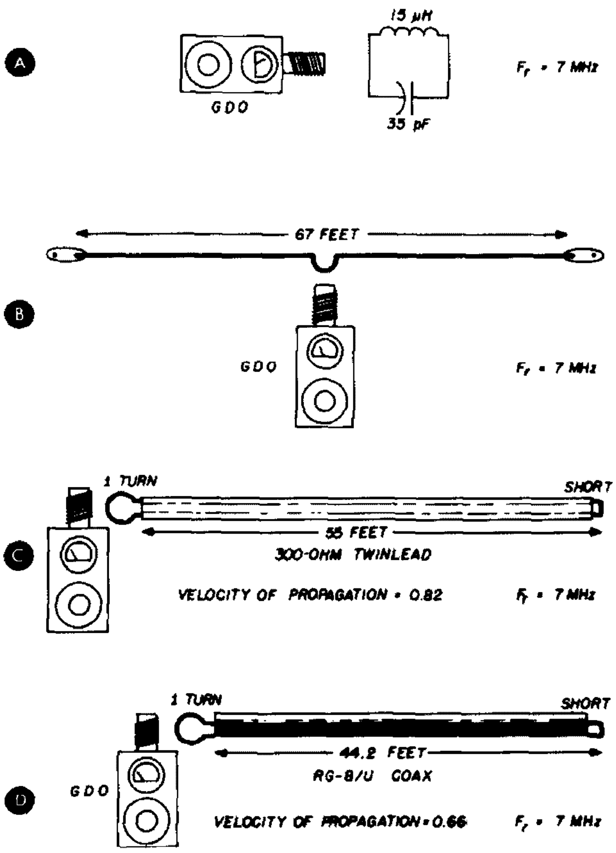


fig. 4. Illustrating the use of the grid-dip oscillator for determining resonant frequency of circuits with lumped and distributed LC constants.

is to determine the lowest frequency of resonance. If this isn't possible, you can grid-dip the line at several frequencies; i. e., at 4, 8, 12, and 16 MHz. A halfwave antenna has a "harmonic family" of resonances, roughly speaking, so you can expect similar behavior from a similar section of transmission line.

coiled transmission lines

Some hams roll up sections of transmission line for compactness. This is where problems can occur. It's practicable to coil sections of coaxial cable, but this procedure isn't recommended for balances lines, such as twinlead.

I experienced some frustrating results with a gdo coupled to coiled twinlead. Normal response was obtained when the twinlead was unrolled from the spool; but with the twinlead rolled up, gdo response was baffling. I obtained a series of minor and major dips, none of which seemed to produce a sensible pattern. When uncoiled, the twinlead behaved in normal fashion.

This anomaly may help to explain some rather odd reactions by those who have used twinlead for matching sections and, when coiling the line, have run into strange and unexpected results.

Neil Johnson, W2OLU

blower-to-chassis adapter

In many large rf power amplifiers it's not practical to mount the blower directly on the chassis — because of either size or noise. The setup shown in fig. 5 allows the blower to be remotely located. A small tin can with both ends removed is soldered to a piece of printed-circuit board. The printed-circuit board is then mounted to the amplifier chassis with sheet-metal screws. A length of flexible hose is run from the blower to the adapter. The hose can be secured with a hose clamp or tape.

Bruce Clark, K6JYO

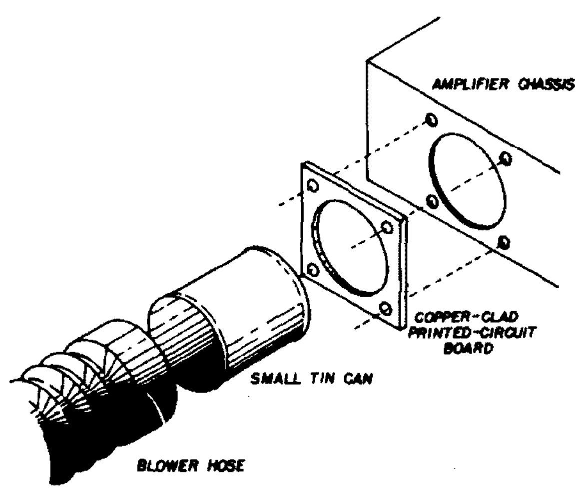
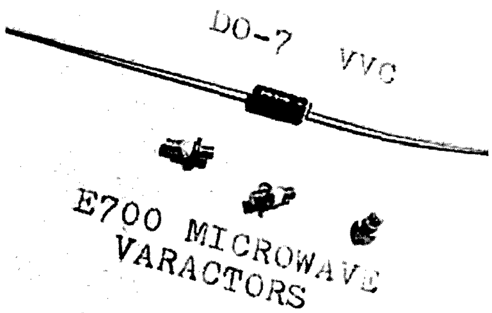


fig. 5. Low-cost blower-to-chassis adapter for high-power amplifiers. Use automobile hose clamp.

new products

amateur radio varactors



A new family of high-Q varactors for amateur radio applications has been announced by the Eastron Corporation. These new varactors offer performance characteristics suitable for operation from lower i-f frequencies up to 450 MHz. Typical uses include automatic frequency control, transceiver incremental tuning, remote control and simplified frequency modulation. These varactors are particularly useful for replacing complex mechanical tuning in compact electronic gear.

Prices range from $1.50 to $6.50 in small quantities. Application notes and data sheets are available from the manufacturer. Write to Eastron Corporation, 25 Locust Street, Haverhill, Massachusetts 01830, or use *check-off* on page 94.

national receiver

National Radio Company's new HRO-600 communications receiver is a high-performance, solid-state receiving system which permits a heretofore unattainable flexibility. The systems concept allows a user to custom-tailor a receiving system to his specific needs at the lowest cost consistent with his requirements. The HRO-600 receiving system consists of a main frame, choice of three frequency control plug-ins, and a large selection of useful accessories.

The main frame contains all receiver signal path circuits from antenna inputs through line and speaker audio outputs. These circuits include an antenna attenuator, slot filter assemblies, frequency converters, i-f amplifiers, i-f filters, a-m and product detectors, and audio amplifiers. It also includes a frequency synthesizer for first mixer injection, a beat-frequency oscillator, and a 115-230 volt 47-420 Hz power supply.

When the main frame is augmented by one of the frequency-control plugins, the receiver is capable of operating at any frequency between 10 kHz and 30 MHz in the following reception modes: a-m, cw, ssb, fsk and fax. Fsk operation also requires an accessory plug-in fsk converter or external audio equipment. Fax reception requires external audio equipment.

Other accessories available for the NRCI HRO-600 include a noise blanker, diversity combiner, remote-control system, dc power supply, vlf/mf/hf preselector, and independent sideband adapter.

For more information on this new receiver, write to National Radio Company, Inc., 111 Washington Street, Melrose, Massachusetts 02176, or use *check-off* on page 94.

hallicrafter fm two-way radios

The Hallicrafters Company has announced two new two-way fm radios that cover the amateur two-meter band as well as the 150-MHz business band; total frequency coverage is from 132 to 174 MHz.

The *Porta-Command* PC-210 is a ten-watt unit which, when combined with quick-change accessories, permits the operator to switch from portable to under-the-dash mobile or base-station operation. Output is from two to ten watts, selectable; hum and noise level is −50 dB minimum; spurious outputs are down 53 dB at 10 watts (46 dB at two watts); power supply voltage is 12 volts dc nominal. The receiver in the PC-210 is a crystal-controlled dual-conversion superhet that features sensitivity of 0.5 μV for 20-dB quieting; spurious and image response is −65 dB minimum; audio output is 1 watt into an 8-ohm speaker. Accessories include battery pack, continuous-tone squelch, ac power tray, battery charger, and back-pack or shoulder carrying case.

The *Porta-Command* PC-230 is a 30-watt solid-state *radio* that weighs only five pounds and takes up less than 250 cubic inches of space. The unit provides up to 12 channels across 1 MHz with no power loss and is instantly adaptable with accessories to mobile, base-station or manpack operation. Power output of the PC-230 is 30 watts; current drain is less than 15 mA on standby, 5.5 amps on transmit; designed for continuous transmit, 100% duty cycle; 132 to 174 MHz frequency coverage. The receiver features sensitivity of less than 0.5 μV for 20-dB quieting; hum and noise are better than 50 dB down from rated output. Accessories include ac power tray, continuous-tone squelch, mobile mounting rack and various antennas.

Illustrated brochures that describe each of these fm sets are available from The Hallicrafters Company, FM 2-Way Department, PR, 600 Hicks Road, Rolling Meadows, Illinois 60008, or use *check-off* on page 94.

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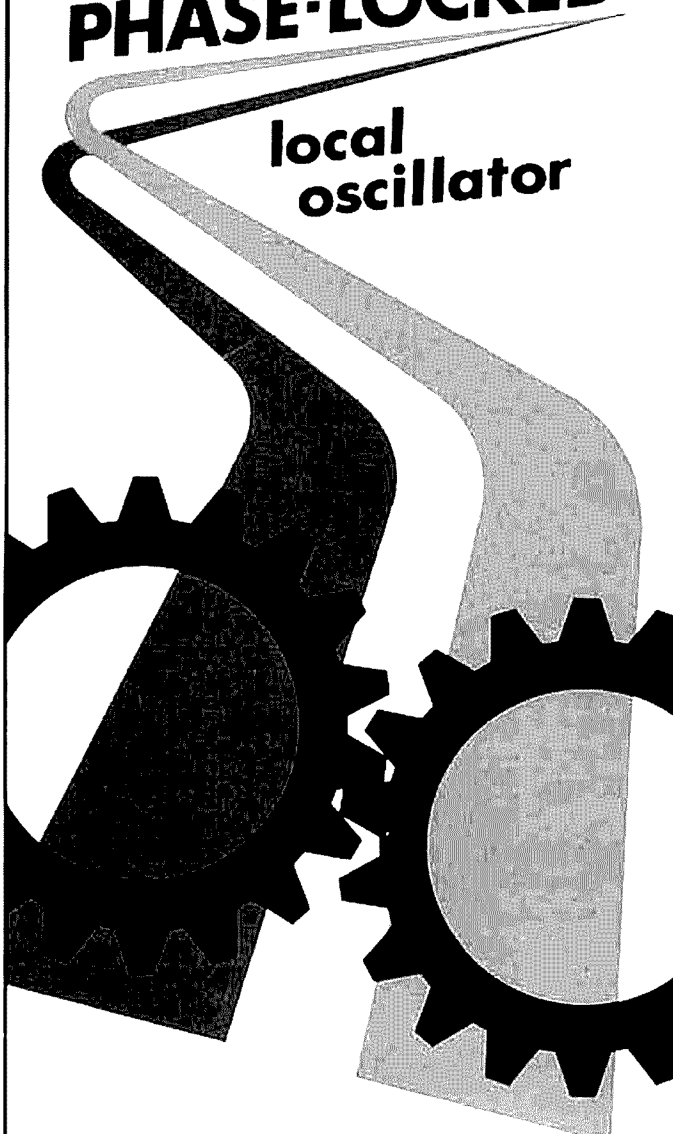
MARCH, 1971

a PHASE-LOCKED local oscillator



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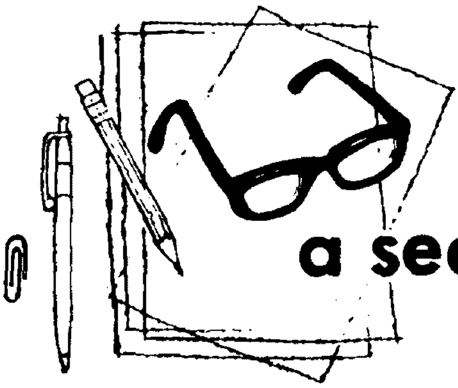


a second look by jim fisk

Within the next 20 years the world's power needs may be supplied by huge solar power stations in synchronous orbit around the earth. Diminishing fossil-fuel reserves, doubts about the practicality of nuclear power and problems of air and water pollution have led scientists to develop a plan for a system of orbiting solar-power stations.

The satellites would collect solar energy with huge arrays of solar cells, and convert the resulting dc into microwave energy for transmission to the earth. On the earth large rectifier-antennas (rectennas) would convert the received microwave energy back to direct current for distribution to the user.

Each power satellite in the system would consist of a solar array about five-miles square, a parabolic transmitting antenna 1.5 miles in diameter and 10,000 microwave transmitting tubes, with an output of 1000 kW each. The transmitting antenna and solar array would be connected with a super-conducting cable 2 miles long. Total power of the proposed system is 100-million kilowatts.

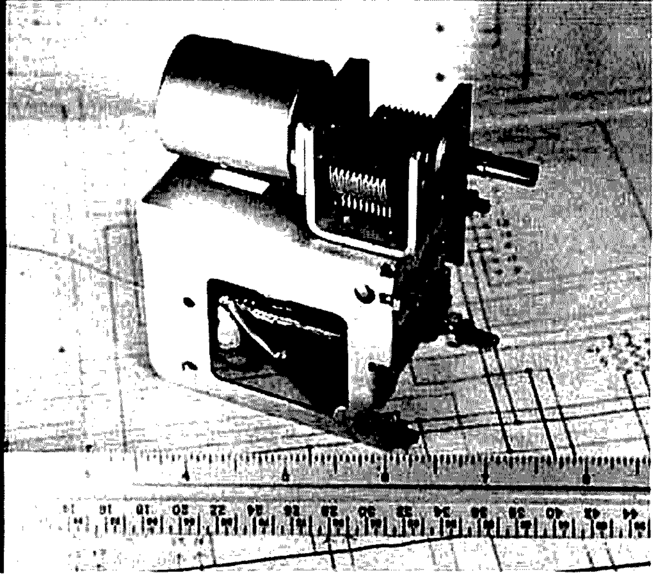
A large rectenna on the earth, several miles square, would convert the received microwave energy back into dc power for distribution. Each element of the large receiving array would consist of a half-wave dipole with its own four-diode bridge. Since the individual dipole-diode combinations (dipodes) are independent, the overall beam pattern is essentially nondirectional, so it eliminates the necessity for precise aiming at the satellite.

An operating frequency of 3000 MHz was chosen to minimize the effects of atmospheric attenuation. Higher frequencies would be attenuated by the trophoshere while lower frequencies suffer from ionospheric attenuation. In addition, in complete cloud cover and moderate rainfall attenuation is lowest around 3000 MHz.

Many people, when first hearing of this plan, imagine that the microwave beam would virtually be a death ray that could cause great damage. Actually, with the proposed 100-million kW system, power density within the beam would only be 10 mW per square centimeter near the earth. This is the official Air Force maximum exposure level, and is a conservative limit with a built-in buffer factor, so it should not be hazardous for living things.

Although 20 years is a reasonable estimate for getting the satellite solar power system into operation, some scientists predict that the system could be working in 10 years if the government were willing to initiate a crash program similar to the nation's moon-landing program. In any event, this is the kind of planning we need to solve the increased demands for electric power in the future.

Jim Fisk, W1DTY
editor



phase-locked local oscillator

A practical solid-state phase-locked local oscillator that covers the range from 14 to 50 MHz

J. A. Koehler, VE5FP, 2 Sullivan Street, Saskatoon, Saskatchewan, Canada

This oscillator was built to satisfy the requirement for the first local oscillator in a receiver. The design called for a number of stable frequencies separated by exactly 1 MHz over as wide a frequency range as possible — at least 30 MHz. One obvious solution would have been to use separate crystals for each frequency. However, this was not economical because of the large number of frequencies. An alternative would have been to build a frequency synthesizer using somewhat fewer crystals (twelve in this case), and produce the required outputs by mixing. I have gone this route before and it is extremely difficult to get all the frequencies precisely where required. The problem of suppressing unwanted mixing products is also formidable. For these reasons I decided to build a variable-frequency oscillator which covered the desired frequency range and which could be phase locked to the harmonics of a 1.0-MHz crystal-controlled oscillator.

Although phase-lock systems are not uncommon in commercial equipment, there has not been a great deal written about them in the ham literature; probably because a rigorous analysis of the process is mathematically complex. However, it is relatively easy to understand their operation and to build systems which work.

The phase-lock system is basically a servo system where the difference between the variable-oscillator frequency and a *reference* frequency produces an

error signal which is used to alter the frequency of the variable oscillator. The process is similar to automatic frequency control (afc) except for one major difference which makes it far superior.

phase-lock operation

A conventional afc system is shown in block form in **fig. 1**. The discriminator produces an output which is proportional to the difference between the frequency of the voltage controlled oscillator (vco) and a reference frequency (the center frequency of the discriminator). The system is designed so that the overall feedback is negative. For example, if the discriminator output is zero volts for the desired frequency, positive for higher frequencies, and negative for lower frequencies, and if the vco moved up in frequency for positive control voltages and down for negative control voltages, then the amplifier must invert as well as provide gain. Then, if the vco drifted up in frequency for some reason, the discriminator output would become positive; the amplifier would amplify this voltage and would make it negative. This negative voltage, applied to the vco, would cause its frequency to decrease, thus counteracting the drift. The problem

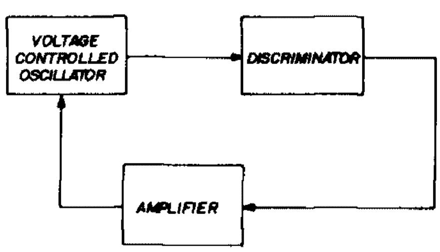


fig. 1. Basic parts of the automatic frequency control circuit.

with an afc system is that frequency drift cannot be compensated for exactly; there is *always* some frequency error.

This can be proved by the following argument. Suppose the vco were originally set exactly at the desired frequency. And furthermore, at that frequency, the discriminator output were exactly zero volts. Therefore, the control voltage to the vco would also be exactly zero volts. Now suppose that the vco drifted slightly upward in frequency. To bring it back to the original frequency requires a negative control voltage. This control voltage could only be produced if the discriminator output were somewhat more positive than zero; i. e., the vco frequency could not be exactly its original value.

This requirement for an error signal is characteristic of any servo system. The error can be reduced by tightening the control loop; in the case of afc this is done by increasing the gain of the amplifier so the control output would be very sensitive to small frequency changes.

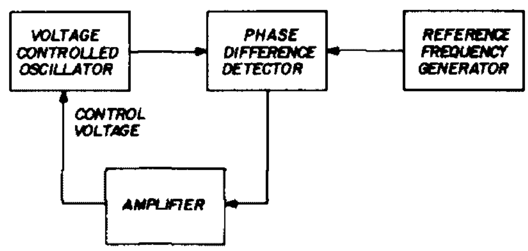


fig. 2. Block diagram of the basic phase-lock system.

However, the frequency error cannot be eliminated entirely. In practice, the vco would be designed to be very stable so that frequency drift would be small. The afc system would then reduce drift even further.

A phase-locked system (pls) can produce an output frequency which is exactly the same as the reference frequency. This is done by deriving the error signal not from a frequency difference but from a phase difference. The block diagram of a pls is shown in **fig. 2**. Note that the afc discriminator has been replaced by a phase-difference detector. A reference frequency oscillator is also required, the discriminator performed this function in the afc system.

Fig. 3 shows a hypothetical time sequence of two input voltages (one at the reference frequency and the other at the vco frequency) and the output of the phase-difference detector. Initially the two voltages are in phase (the phase difference was zero) so the output of the phase detector is zero. However, because

the two frequencies are different, the phase *difference increases steadily with* time until it reaches 360°. At that point the output of the phase difference drops back to zero since 360° phase difference is equal to zero phase difference.

Now suppose the output of the phase detector were connected to the vco control line. The ramp voltage would sweep the vco frequency. For the system to work properly, the control-voltage would sweep the vco frequency in a direction so as to decrease the frequency difference. Then, as the vco frequency approached the reference frequency, the rate of change of phase difference would decrease, and the vco would sweep more slowly. When the vco exactly reached the reference frequency, the phase difference between the two would be constant in time, so the vco frequency sweep would stop; the vco frequency would be phase-locked to the reference frequency.

The error signal in this loop is due to the phase difference between the vco and the reference frequency while they are locked together. This phase difference can be made arbitrarily small, just as the frequency difference in an afc can be reduced, but it cannot be entirely eliminated. However, for purposes of frequency control it is immaterial because the frequency error will be zero regardless of the phase error. If the vco frequency tended to drift, the phase detector would produce an output voltage which would

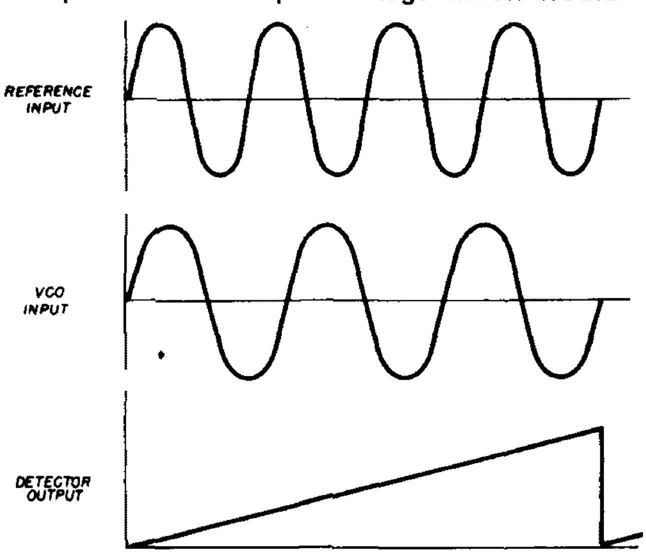


fig. 3. Waveforms in the phase-difference detector.

keep the vco locked to the reference frequency. After the drift, the phase difference between the two oscillators would, of course, be different from its original value, but the two frequencies would still be identical.

In practice, it is difficult to build a phase-difference detector with characteristics exactly as outlined above, and the devices used are called phase-sensitive detectors. The output of phase-sensitive detectors are usually only approximately

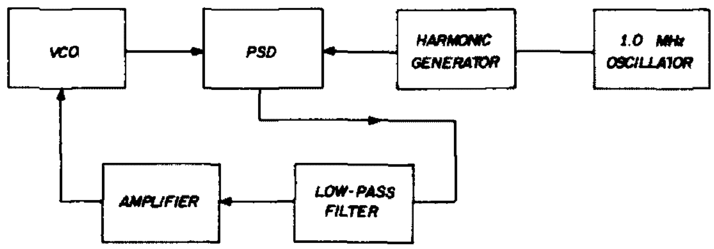


fig. 4. Proposed phase-locked oscillator for receiver local oscillator.

proportional to phase difference, and then only over a restricted range of phase difference. Also, there will also be a low-pass filter of some sort following the detector.

The lock-in range of the system, the range of vco frequency variation over which it can be locked onto the reference frequency, depends in a complicated way upon total loop gain and low-pass filter band-width. If you are interested in pursuing the topic, there are a surprising number of papers in technical journals on the subject. One of the more readable explanations I have seen appeared in the *Proceedings of the IRE*[1].

practical phase-locked oscillator

For my requirements, the vco had to have as wide tuning range as possible. Fair Radio has been selling a surplus tuning unit with a ganged capacitor and slug-tuned coil that covers the frequency range from 14 to 50 MHz, and I bought one of these.* I wanted to use this mechanism as the basis for a variable oscillator for this frequency range and

*$1.25 from the Fair Radio Sales Company, Post Office Box 1105, Lima, Ohio 45802. Order RT-45 *tuner. Shipping weight: 1 pound.*

phase lock it to the harmonics of a 1-MHz oscillator.

The system is shown in fig. 4. The surplus tuning unit was made into a Hartley oscillator using an fet. With no padding capacitor, the frequency range extended from about 15 to 80 MHz — this was more than adequate, and a padding capacitor was added to reduce the range somewhat, and to lower the low-frequency limit to a bit lower than 14 MHz. Frequency stability was quite

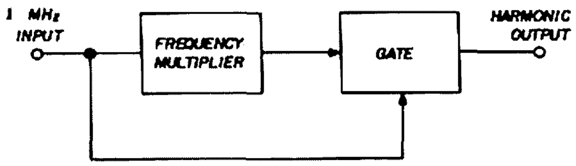


fig. 5. Harmonic generator.

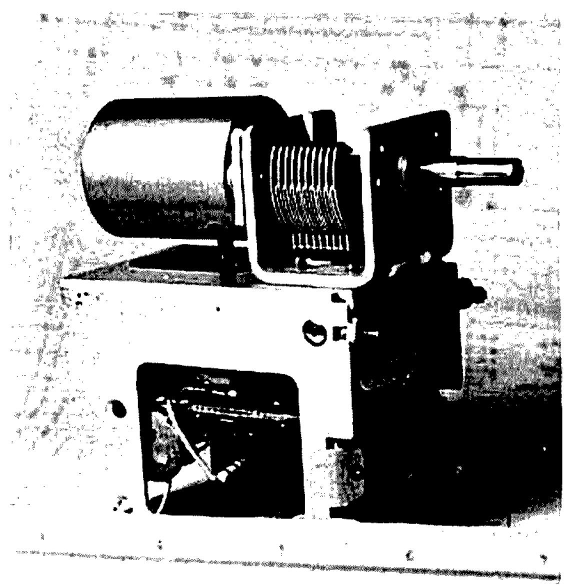
good. Normally the oscillator should be as stable as possible so that it does not readily drift out of the lock-in range. A varicap across the tank circuit was to be used as the frequency controlling element.

The 1-MHz crystal-controlled oscillator also used an fet. The harmonic generator was more complicated than the type usually seen in the amateur magazines. A phase-look system requires that some power must be supplied to the reference frequency input to the phase detector at all frequencies where phase lock is required. To lock at 50 MHz, for example, requires appreciable amplitude of the 50th harmonic of the 1-MHz oscillator. Harmonics were originally generated by squaring the 1-MHz waveform and applying it to a diode — and later, other non-linear devices.

However, after a few experiments, it was apparent that the amplitudes of the higher harmonics generated in this way were too small. Step-recovery diodes probably would have been excellent; however, I didn't have any, so I built the harmonic generator outlined in fig. 5.

The output of the multiplier was keyed on and off at a 1-MHz rate by the gate. Assume that the frequency multiplication factor were ten and that it was somewhat non-linear; the output of the multiplier would then contain energy not only at 10 MHz, but also its harmonics, 20, 30, 40 MHz, etc. The gate can be thought of as a modulator which overmodulates. Consider for example the signal from the multiplier at 10 MHz. This would be modulated at a 1-MHz rate so the output of the gate would contain the carrier at 10 MHz as well as sidebands 1 MHz to each side of it (9 MHz and 11 MHz).

Because of overmodulation, there would also be other sidebands 2, 3, 4, and 5 MHz on either side of 10 MHz. The output of the gate would contain all the carriers (spaced at 10 MHz intervals) plus all the sidebands (spaced at 1 MHz



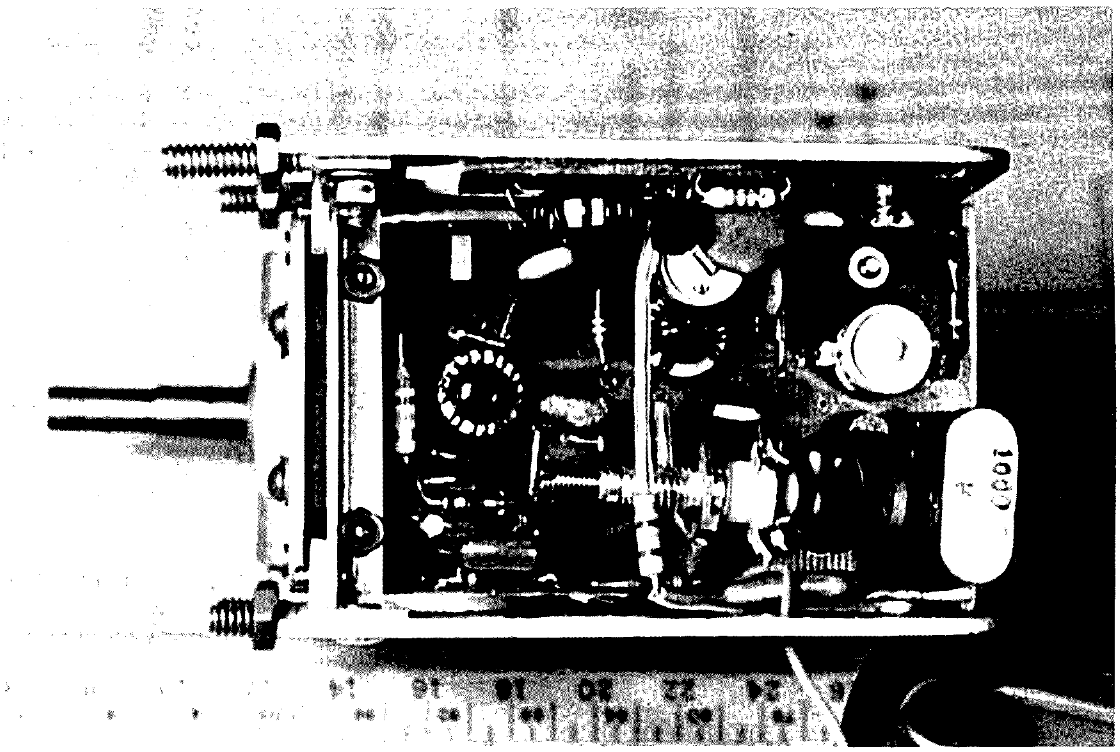
Miniature phase-locked local oscillator.

intervals) between the carriers. This type of harmonic generator can be very efficient. The simple version I built produced sufficiently strong harmonics up to 70 or 80 MHz.

Since any mixer is, in fact, a phase-sensitive detector of sorts, my phase detector was a simple diode double-balanced mixer. I didn't have any hot-carrier diodes, so four 1N270s were selected by matching them on a curve tracer. The low-pass filter was an RC type, and the amplifier was a μA741 integrated circuit.

Unfortunately, when the system was built, it would not work properly over the entire range of the vco. Despite careful diode matching, the double-balanced mixer was not well enough been to provide some amplitude limiting of the vco output. However, the system was beginning to get too complicated, and the whole approach was abandoned.

After considerable experimentation,



Complete phase-locked oscillator circuit is built on small board that is mounted in the surplus LC tuner.

balanced. The output of the mixer must be coupled directly to the amplifier input. Ideally, the output of the mixer should be at zero volts dc when the two inputs are not harmonically related. However, due to diode unbalance, there will always be a small dc component as well as the desired signal which offsets the amplifier input.

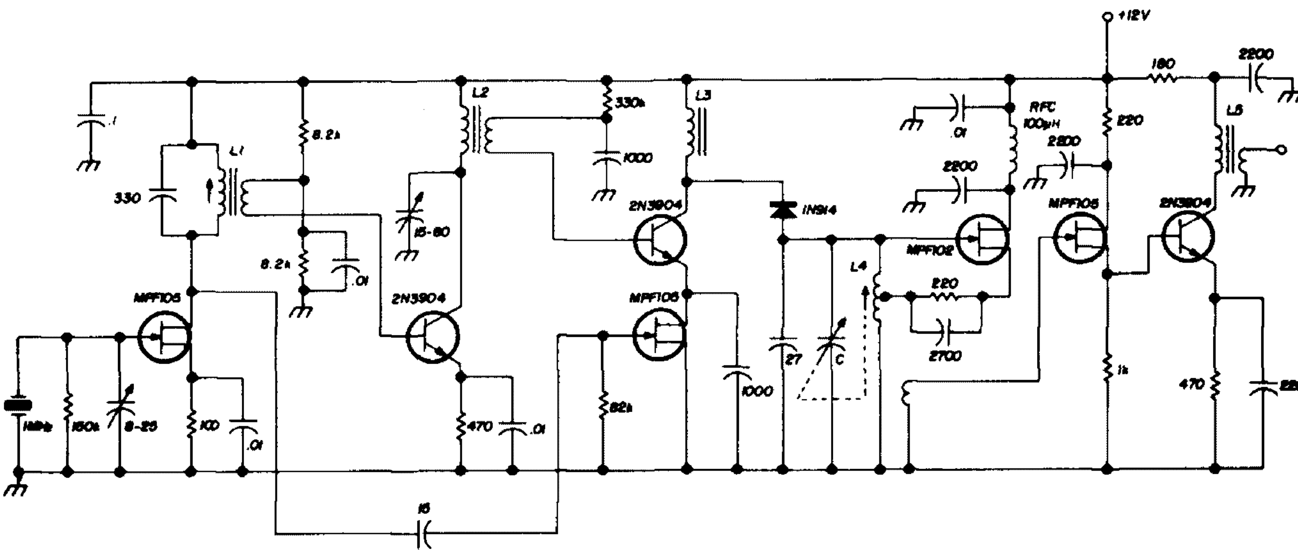
The problem was that the offset depended on vco amplitude. The system could, for example, be adjusted to work near 14 MHz by compensating for dc offset. However, when the vco was tuned to higher frequencies, the output amplitude changed, causing the dc offset to change and saturate the amplifier. If the gain of the amplifier were reduced so that it did not saturate anywhere over the vco range, then the lock-in range became unacceptably small at the higher frequencies. A possible solution might have

the final solution required very few components. The oscillator fet itself was used as a mixer by coupling the harmonic generator output into the oscillator tank circuit with a varicap. The signal at the gate of the fet contained the oscillation frequency as well as all the 1-MHz harmonics. The instantaneous voltage at the gate depended on the phase difference between the oscillation frequency and the harmonics. Because the varicap was connected to the gate, the oscillation frequency therefore depended on this phase difference — thus satisfying the requirement of a phase-locked loop.

This system worked very nicely when I used a 1N914, a silicon planar switching diode, for the varicap. Normal varicaps with capacitance of 10 pF at 4 volts resulted in lock-in ranges which were too large, and which overlapped. With the 1N914, the lock-in range was about 100

kHz at 14 MHz and somewhat lower at the higher frequencies.

The final circuit is shown in fig. 6. The oscillator transistor was mounted on small terminal strips next to the coil. The factory, and it is anticipated that when the system is incorporated into the receiver, it will not be necessary because considerable wideband noise will be generated when the system is not locked.



C1, L4 Part of surplus tuner, Fair Radio RT-45 (see text)

L1 100 — 125 μH (J. W. Miller 42A104CBI) with 20-turn link

L2 25 turns no. 28 on Phillips 2P65347 toroid core, 7-turn link

L3 18 turns no. 28 on Phillips 2P65347 toroid

L5 15 turns no. 28 on Phillips 2P65347 toroid, 4-turn link

Phillips toroids are 0.375 inch in diameter. Ferroxcube 266CT125 toroids made of 4C4 material are identical to the Phillips units; Indiana General type CF102-Q1 are virtually the same.

fig. 6. Circuit of the practical phase-locked oscillator that tunes from 14 to 50 MHz.

1-MHz oscillator and harmonic generator was built on a printed-circuit board which fit nicely inside the framework of the surplus capacitor-coil framework as shown in the photograph. The smaller board on the side of the frame is the output buffer amplifier.

The frequency multiple in the harmonic generator need not be ten but can be any multiple. In the final version, it was sixteen.

The only remaining detail was to determine whether the vco was free running or whether it was phase locked at any point on the dial. This was done by connecting a diode detector and audio amplifier in parallel with the output. As the vco was tuned through its range, it was possible to hear it snap in and out of lock.

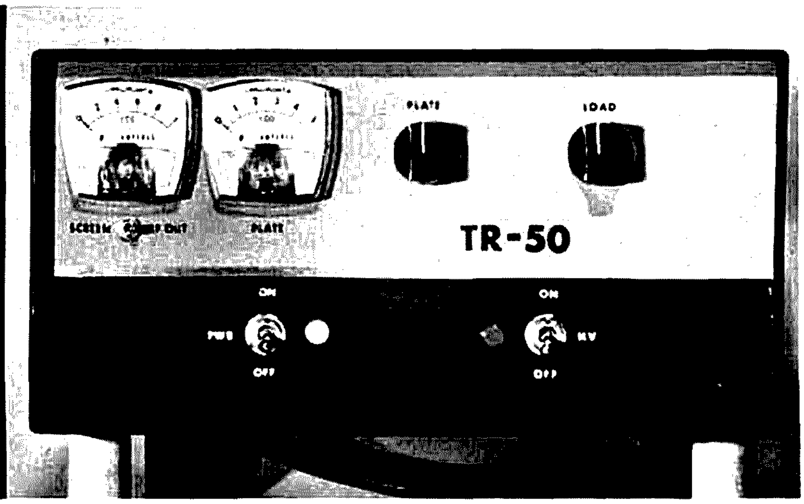
This method is not particularly satis-

As an interesting side effect I found that it was possible to lock the vco on to harmonically related frequencies other than the 1-MHz points. For example, the vco could be locked to exactly 14.3333 MHz because the mixer detected the phase difference between the third harmonic of the vco at 43 MHz and the 43rd harmonic of the 1-MHz oscillator with sufficient amplitude to cause locking. However, the lock-in range for these frequencies is quite small and there is no difficulty in determining whether or not the output frequency was on one of these points or on the desired frequency.

reference

1. H. T. McAleer, "A New Look at the Phase-Lock Oscillator," *Proceedings of the IRE*, June, 1959, p. 1137.

ham radio



TR-50 customized six-meter transverter

A compact 300-watt transverter featuring high efficiency and a low-noise mosfet front end

Louis E. Savoie, K1RAK, 29 Hillsdale Road, Holbrook, Massachusetts 02343

The **50-MHz transverter** described in this article was built to satisfy a specific purpose: to provide a medium power, tabletop transverter which was compatible, both electronically and in styling, to the Drake TR-3. A secondary object was cost. The result is a 300-watt ssb transverter that uses a low-noise mosfet receiving converter, and includes built-in solid-state power supplies. Efficiency of the output stage is 64%.

cabinet

My first order of business was to select a suitable enclosure for the unit. The Drake TR-6 cabinet matches the TR-3 in styling and offers the advantage of not having additional cutouts as do other cabinets of the TR-4 line. Cabinet measurements indicated a 10-1/2 x 13-1/4-inch chassis and 10-1/2 x 5-1/4-inch front panel were optimum. Considering the weight of the power-supply components, 0.047 steel was chosen for the chassis to provide strength and rigidity.

The chassis and front panel were purchased from a local metal shop for $9. This included bending and corner welding. The chassis, front panel and cabinet were then mated, and the chassis drilled

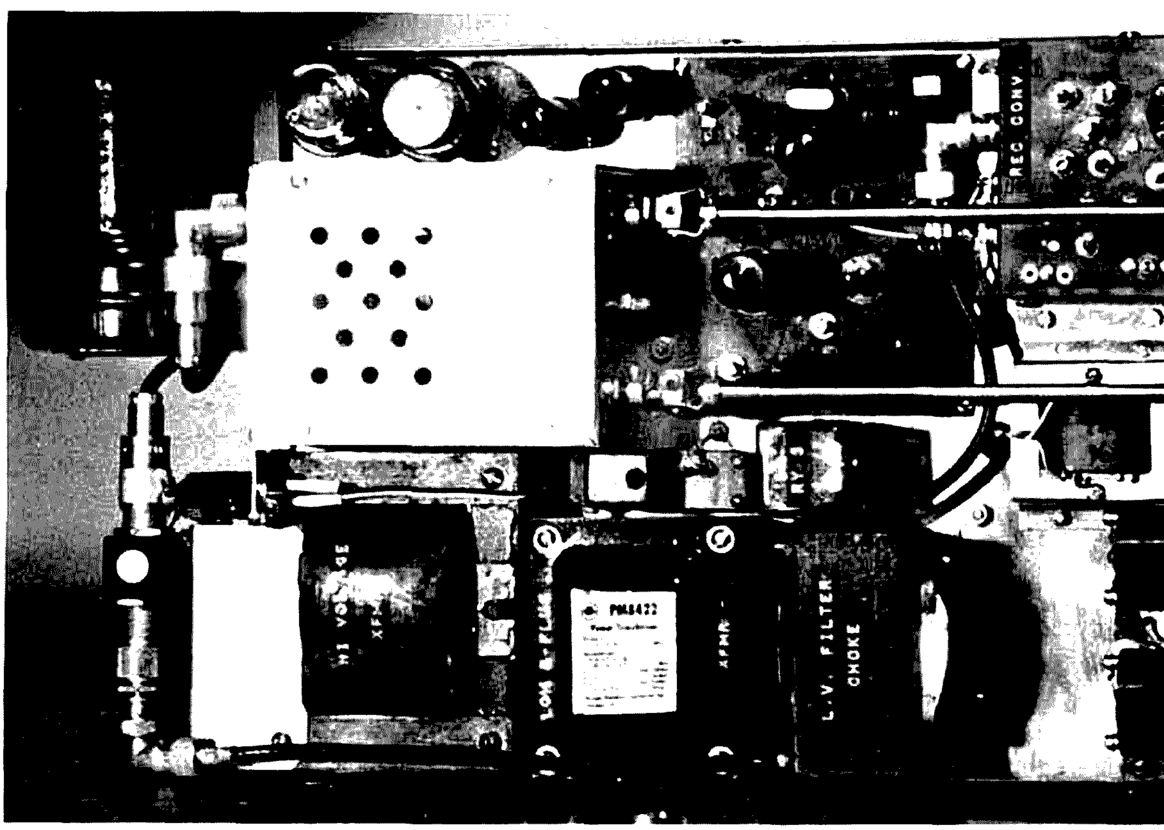
for the captive hardware I wanted to use to hold the system together. The captive nuts should be riveted in place to provide a smooth, flush outside surface.

circuit

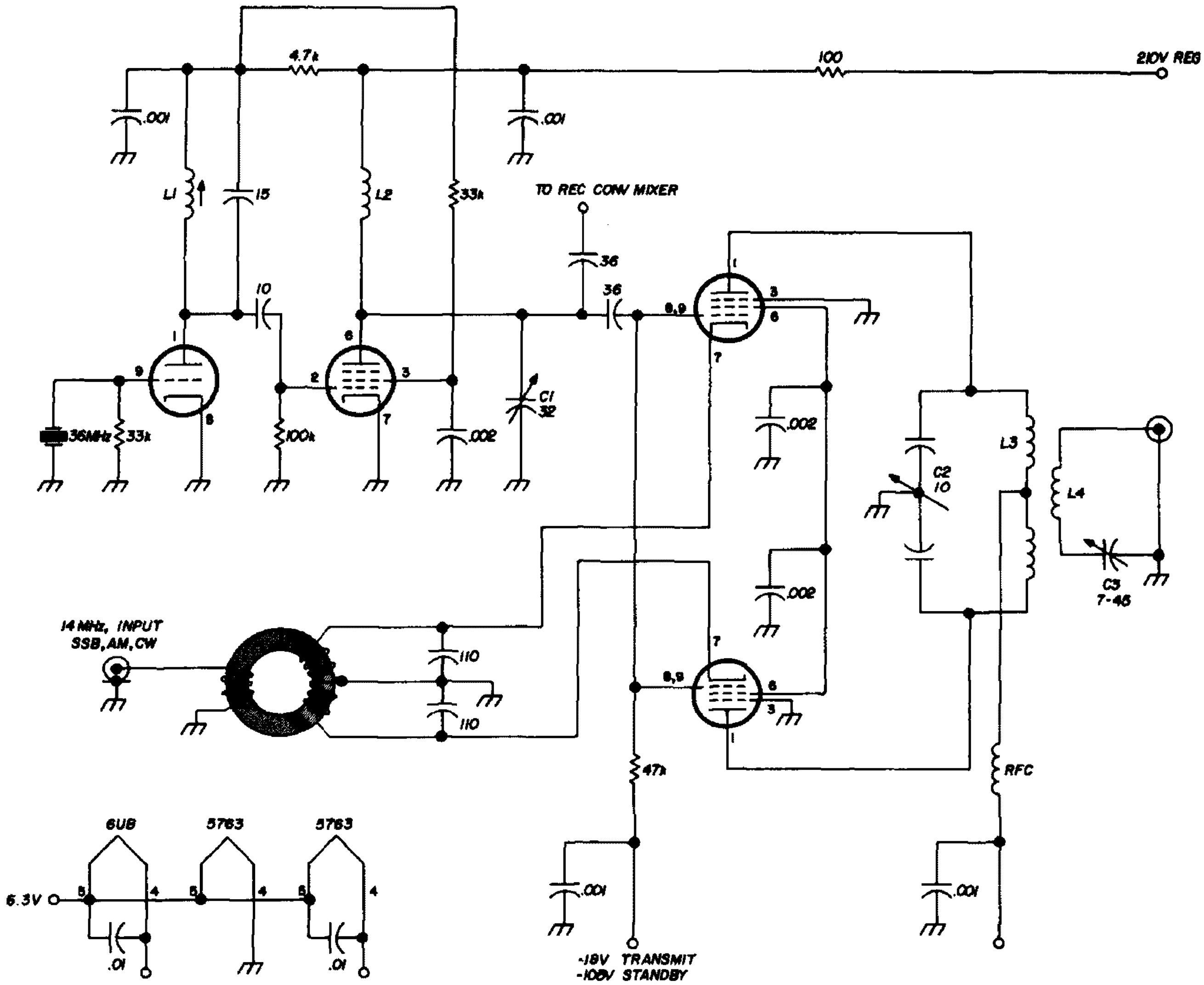
In the next phase of the design I considered both the mechanical and electrical requirements of the transverter. I have found it useful to first decide upon the circuit and power-tube requirements, and then search the literature for commercial transformers that will do the best possible job. Since I wanted the design to be as close to the state-of-the-art as possible, I decided on printed-circuit construction; solid-state would be used where feasible.

A recent article by K2ISP[1] described printed-circuit transmitting mixers for both six and two meters. Local experience with the two-meter unit suggested similar possibilities for the six-meter mixer. K2ISP's mixer features a conventional 6U8 buffer/amplifier that feeds a push-pull 5763 mixer/driver stage; the low-frequency ssb signal is injected at the cathodes of the 5763s. Although the original circuit was designed for a 43-MHz local oscillator and 7-MHz ssb, it seemed reasonably easy to convert to the more popular 36-MHz local oscillator, 14-MHz ssb system (fig. 1).

The printed-circuit board used in the K2ISP transverter, shown in fig. 3, can be used in the TR-50 without modifica-



Top view of the TR-50. The coaxial cable on the bottom goes to the receiving converter.



C1 32-pF variable (E. F. Johnson 160-110)

C2 10-pF butterfly (E. F. Johnson 160-211)

C3 7 — 45 pF ceramic trimmer

L1 12 turns no. 24, close wound on 1/4" slug-tuned form

L2 12 turns no. 20, air wound, 1/4" diameter, 1/2" long

L3, L4 7 turns no. 20, air wound, 1/2" diameter, 9/32" long

L5 4 turns no. 20, air wound, 1/2" diameter, 5/32" long

RFC Ohmite Z-50

Xtal 36-MHz third overtone

fig. 1. Six-meter transmitting mixer uses 14-MHz ssb input.

tion — although a few component values must be changed to accommodate the new mixing frequencies. The oscillator and buffer coils were re-designed to cover 36 MHz, and the input toroid was wound for 14-MHz injection.

After all the components were mounted on the printed-circuit board, the unit was tested on the bench. Using a TR-3 for injection, through a 10-dB pad, the transmitting mixer worked well into a 50-ohm load. A Heath Monitorscope indicated good quality ssb. Power output was calculated to be slightly more than 1.5 watts — more than enough to drive a power amplifier stage.

power amplifier

The available chassis space dictated a power tube that would combine ef-

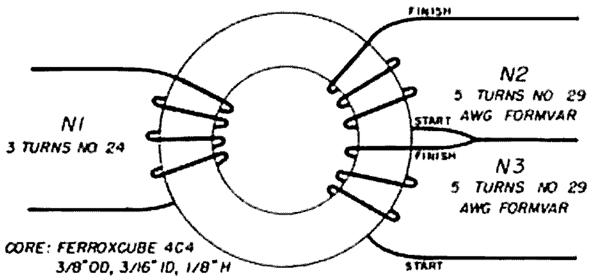


Fig. 2. Toroid transformer for 14-MHz input. Core is Ferroxcube 4C4 3/8" OD, 3/16" ID, 1/8" thick. Winding N1 is 3 turns no. 24 enameled: Winding N2 and N3 are each 5 turns no. 29 enameled. Windings N2 and N3 must be wound in same direction. Primary and Secondary winding should use full circumference of the core.

ficiency with small size. The 4X150/4CX250B power tetrodes offer a power-output-to-size ratio that seemed ideal for this application. These tubes require forced-air cooling, but commercial squirrel-cage blowers are available in small sizes and at low cost. A schematic of the six-meter power amplifier is shown in fig. 4.

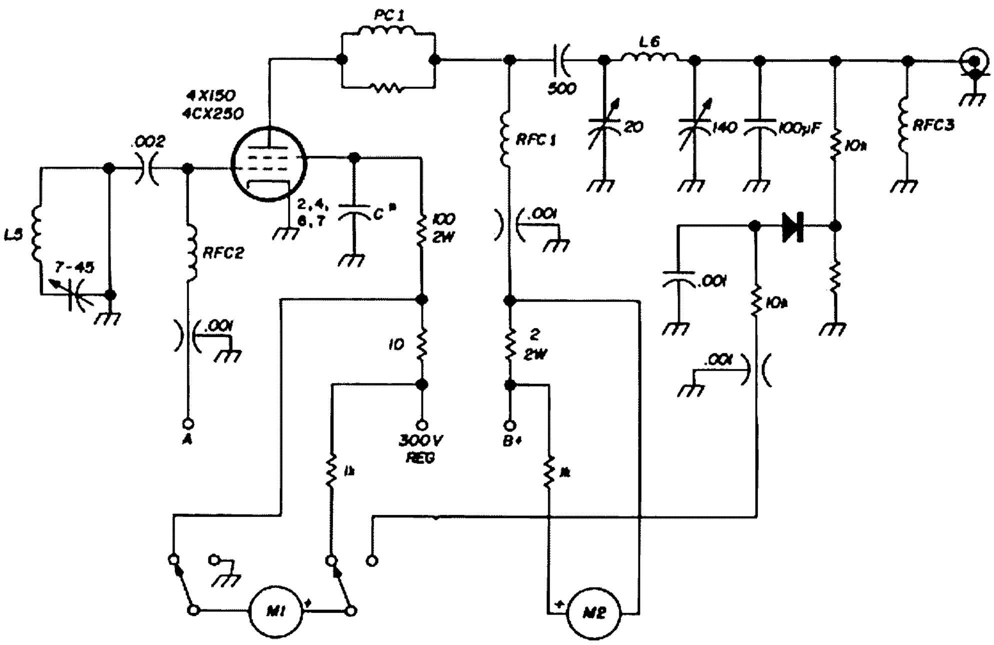
Rough sketches indicated that the pi-network coil, variable capacitors, plate choke and bypasses would easily fit into a 3x4x5-inch aluminum utility box. This made a suitable rf compartment which would fit into the vertical space between the cabinet and the top of the chassis.

The grid compartment box was made from 0.047-inch steel and mounted upside down under the chassis. By mounting this box against the back edge of the chassis, and including a rubber-gasketed cover, a plenum chamber was created. Cooling air is provided by a Dayton 21-cfm blower mounted to the chassis and sealed with rubber gaskets. A 1½-inch diameter hole through the back of the chassis and the grid compartment provides for air flow.

All power and filament connections are fed to the grid compartment through, .001 μF feedthrough capacitors. Coupling

fig. 3. Full-size printed-circuit board for the transmitting mixer.

from the mixer output link is fed to the 4CX250B grid through a ceramic feed-through connector. I believe this setup provides the input-output isolation which resulted in an exceptionally stable power vided power for the miniature coaxial relay and 15-volt zener-regulated receiving-converter power supply.* Power supply circuits are shown in figs. **9, 10** and **11.**



C1 Eimac SK600 socket bypass capacitor

L5 Link on transverter board (L4 in fig. 1)

L6, 5 turns 1/8" copper tubing, 1" diameter

PC1 3 turns no. 14 around 47-ohm, 2-watt resistor

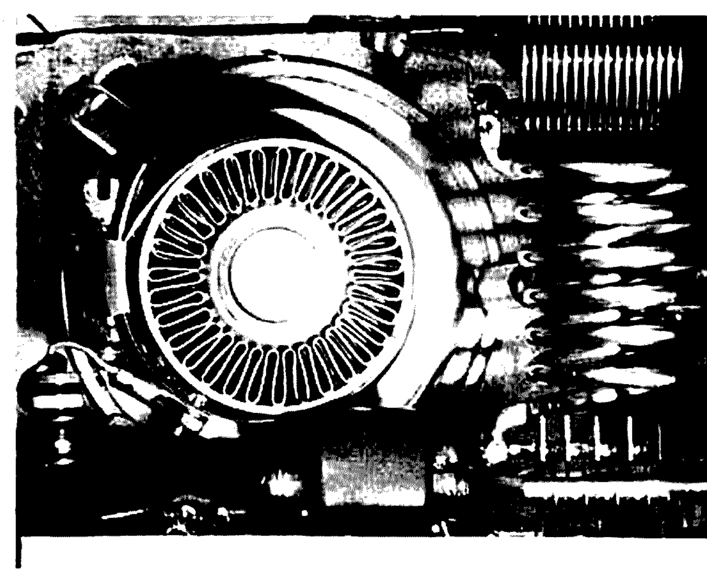
RFC1,2,3 Ohmite Z-50

fig. 4. 4X150 power amplifier stage operates at 64% efficiency and requires no neutralization.

amplifier that needs no neutralization.

I used low-cost surplus transformers and chokes for the various power supplies, although suitable commercial transformers are available. Separate transformers were fused for the high- and low-voltage supplies. A Stancor PS8415 was used for the bias supply, and a small 24-volt electric-train transformer pro-

*Shortly after completing this project, I uncovered a small bonanza. Jim Reeves, WA9HKE, 2207 Columbus Avenue, Anderson, Indiana purchased the closing stock of an electronics manufacturer, and included in this stock were a number of multiple-winding power transformers yielding 1100 to 1200 volts, 300 volts, 130 volts, 6.3 volts and 5 volts. The price is $12 each. These transformers are ideal for transverter projects such as the TR-50; I bought several for future projects.



Power-amplifier compartment. The 4CX250B is mounted to the rear of the compartment. Tube clamp is a stainless-steel heater-hose clamp.

switching and controls

Since the power output of the TR-3 is much higher than that needed for the 5763 mixer circuit, it must be attenuated. The 10-dB pad shown in fig. 5 is somewhat over-designed, since the resistors specified will permit 220 watts dissipation. It is built into a 3x4x5-inch minibox and connected to the transverter input with short length of RG-58/U coaxial cable.

The switching relays consist of one three-pole, double-throw unit (K3), a standard Dow-Key antenna changeover relay (K1), and a miniature surplus coaxial relay (K2). K3 is the main control relay and is actuated by extra contacts in the ssb exciter (fig. 6); it actuates K1 and K2, and grounds the two bias lines. One multiple-pole relay would probably do the job, but these three relays were already in my junkbox.

metering

The meter circuits were designed with two factors in mind: low cost and easy-to-calculate meter shunts for the scales I wanted. By using low-cost imported meters in series with 1000-ohm resistors, all shunts are multiples or submultiples of 10 ohms. For example, with a 1-ohm shunt, this system will read 1 ampere full scale: for 500-mA full scale a 2-ohm shunt is required, and for 100-mA full scale, a 10-ohm shunt.

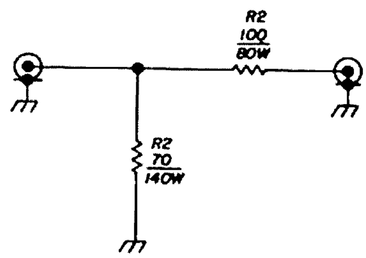
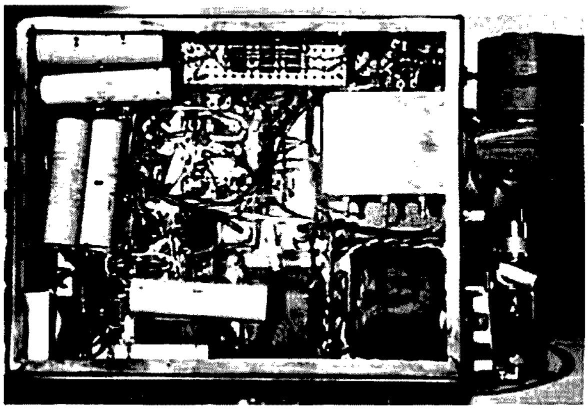
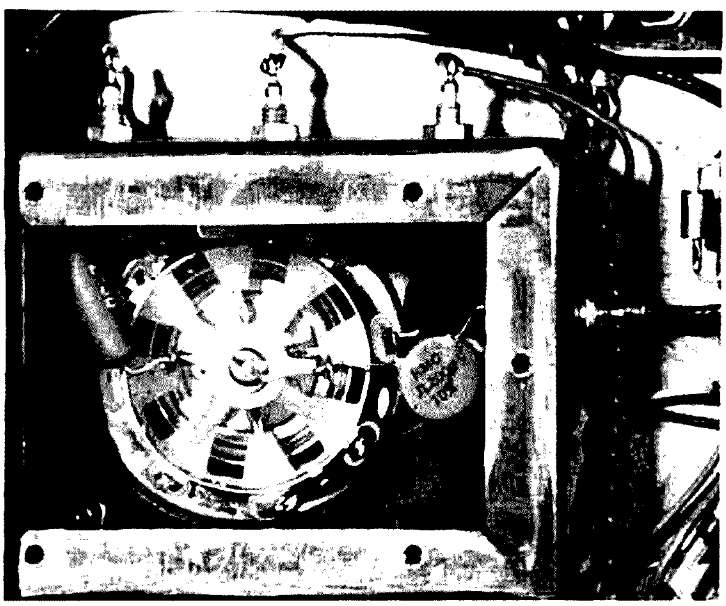


fig. 5. 10-dB pad must be used to lower output of ssb exciter. 100-ohm resistors consist of six parallel-connected 600-ohm 14-watt resistors (Sprague 459E7015). 70-ohm resistor consists of 7 parallel-connected 700-ohm, 14-watt resistors (Sprague 459E6015).



Bottom view of the TR-50. Bias transformer is mounted in lower left corner. Diodes in upper center are the high-voltage bridge rectifiers. BNC fittings in center are from the miniature coax relay.



Power-amplifier grid compartment. All connections are fed through .001 µF feedthrough capacitors.

The meters are used in the plate and screen circuits of the power-amplifier stage; these are the most critical measurements in 4X150 tetrodes. In addition, the screen-current meter is switched to read relative power output. The scale numbers on the original Micronta* meters were erased and replaced with press-on numbers to indicate the desired ranges.

*Micronta meters are $2.98 from Radio Shack.

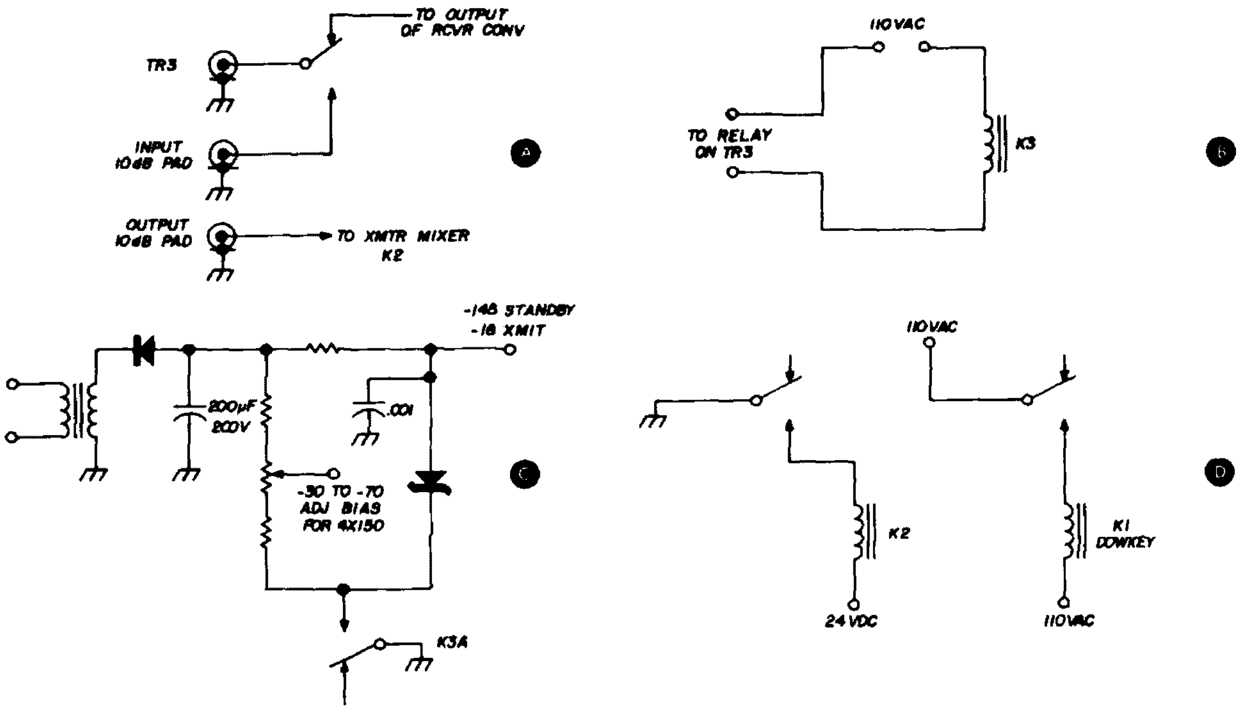


fig. 6. Switching and control circuitry. Relay K3 is a multiple-contact unit actuated by TR relay in the exciter. K2 is a miniature coax relay. K1 is a Dow-Key antenna changeover relay.

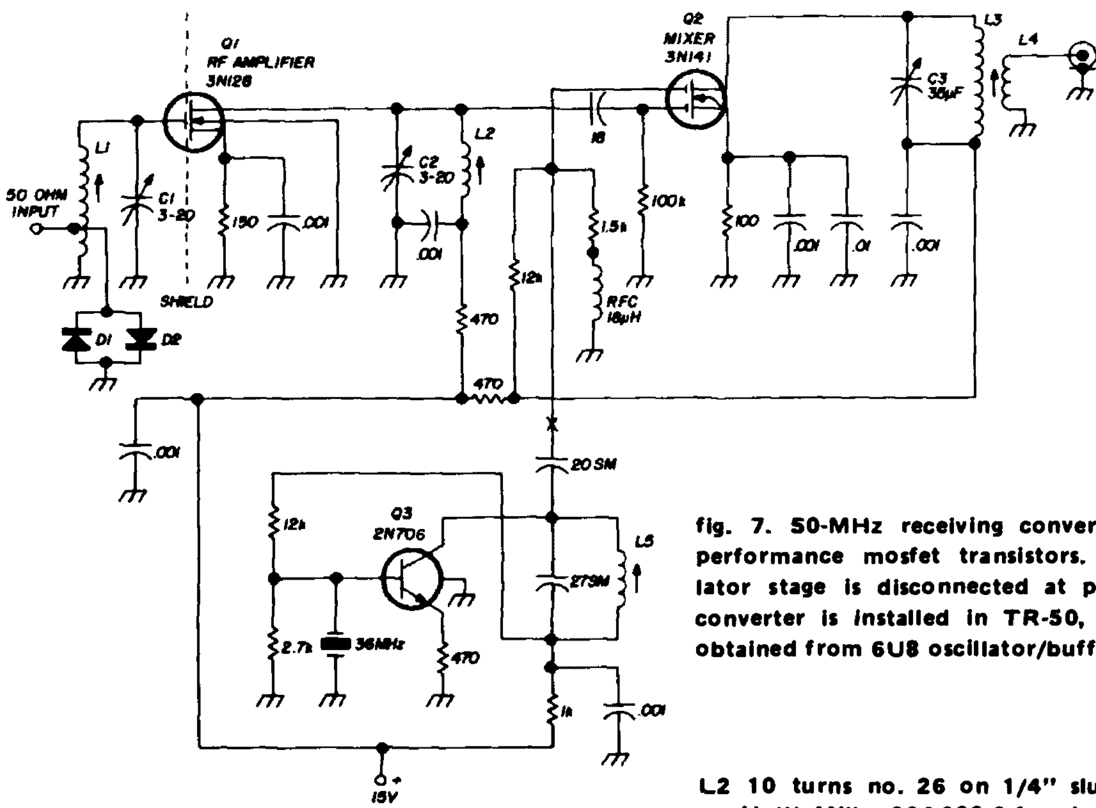


fig. 7. 50-MHz receiving converter uses high performance mosfet transistors. 2N706 oscillator stage is disconnected at point X when converter is installed in TR-50, and injection obtained from 6U8 oscillator/buffer.

C1,C2 3-20 pF air variable

C3 7-35 pF air variable

D1,D2 1N100 or equivalent

L1 10 turns no. 30 on 1/4" slug-tuned form, tap at 3 turns (J. W. Miller 20A000-2 form)

L2 10 turns no. 26 on 1/4" slug-tuned form (J. W. Miller 20A000-2 form)

L3 24 turns no. 26 on 1/4" slug-tuned form, secondary 3 turns no. 26 (J. W. Miller 20A000-2 form)

L4 9 turns no. 26 on 1/4" slug-tuned form (J. W. Miller 20A000-2 form)

RFC 18 µH (J. W. Miller 9310-42)

receiving converter

The receiver converter is based on a design by WB2EGZ originally published in *ham radio.*[2] The modified converter in the TR-50 uses a high-gain single-gate 3N128 mosfet as an rf amplifier; the mixer is a dual-gate 3N141 mosfet. To provide full transceive ability, oscillator injection is taken from the transmitting mixer. This circuit exhibits an excellent noise figure, and has high resistance to cross modulation.

I had originally planned to build the receiving converter into a commercial cast-aluminum box. However, there was not enough space between the top of the chassis and the final-amplifier tuning shafts to accommodate the commercial box. The box I finally used was made from double-sided printed-circuit board. The top and sides of the box were fixed in a holding jig and joined by solder fillets on all interior seams. This chassis is mounted next to the transmitting mixer board for direct capacitance coupling to the output of the 36-MHz oscillator/buffer.

Placing of components in the receiving converter is not critical with the exception of the interstage shield in the 3N128 rf amplifier. The transistor socket should be mounted so the shield can be soldered across the width of the box, isolating the input to a small compartment on one end of the chassis. Ceramic feedthrough connectors are used for oscillator injection and i-f output. A BNC coaxial connector is used at the input.

The converter circuit shown in **fig. 7**

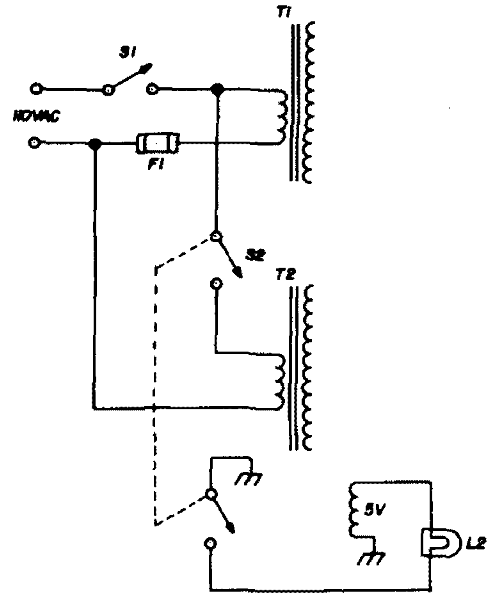


fig. 8. Ac control wiring for the TR-50. Fuse F1 is a 5-amp slo-blow unit.

has a 36-MHz oscillator stage that was included so the converter could be bench tested before putting it into the TR-50. When the receiving converter is installed in the transverter, the oscillator injection line is disconnected from the 20-pF capacitor and connected to the output of the transmitting-mixer buffer. The built-

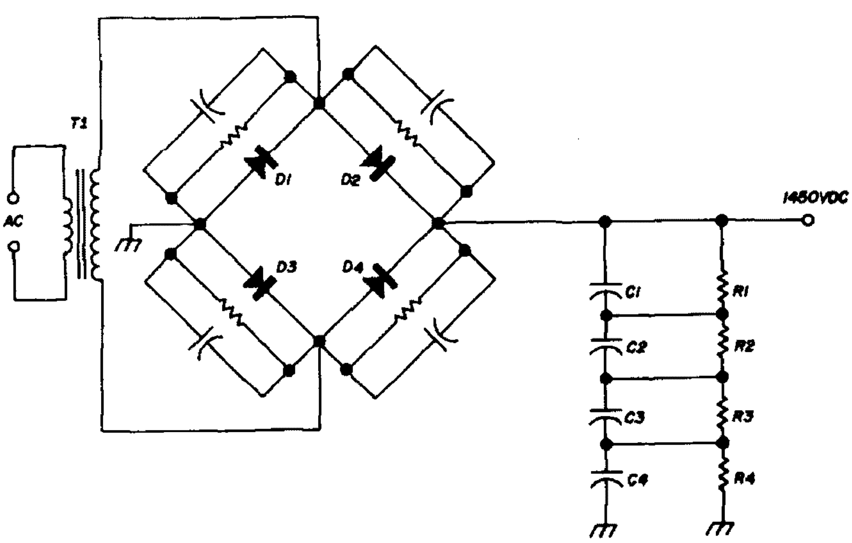


fig. 9. High-voltage power supply for the TR-50.

in receiving-converter oscillator stage is left on the board for future trouble-shooting and maintenance.

alignment and test

After all the wiring and assembly were completed, the unit was given a good shaking out and cleaning to remove all the bits of wire that had become lodged in the crevices. Next, I carefully inspected all the tube sockets, tie points and cable terminations for miswiring, shorts and unsoldered joints. This is time well spent, and it *pays* to do some careful circuit tracing. Do not short change yourself – a simple wiring mistake can become very costly when the power is turned on.

Before applying power for the first time, remove all the tubes from their sockets. Then turn on the power and measure the voltages. The voltages will tend to be on the high side, but will confirm that connections have been made to the right places.

Test the transmitting mixer first. Connect the output of the exciter through the pad to the mixer. Plug the mixer tubes in, along with the OB2 regulators and relay keying circuit. Adjust all tuned circuits to the correct frequency with a grid-dip meter. Temporarily connect a 51-ohm, 2-watt resistor from the output link of the mixer to ground; *do not* install the 4X150 yet.

Tune the ssb exciter for normal output, and using the grid-dipper as a wavemeter, adjust the 6U8 tuned circuits for maximum output to the 5763. Adjust the 5763 tuned circuits for maximum indication at 50 MHz. Be careful not to exceed the tune-up limitations of the exciter.

Turn off the exciter, disconnect power, discharge all the filter capacitors and install the 4X150. Remove the 51-ohm resistor from the mixer output. Connect a 50-ohm non-inductive load to the output of the power amplifier. Apply power and allow a normal warm-up

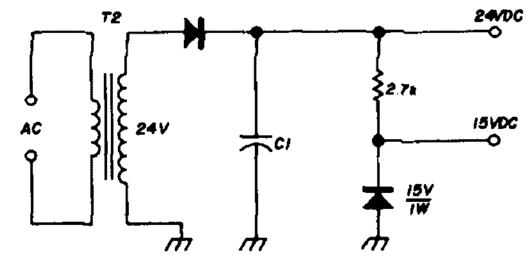


fig. 11. Receiving-converter and relay power supply.

period. Turn the exciter gain control down, key the exciter and adjust the bias on the 4X150 for resting current of 50 mA.

Now turn up the exciter *gain* control and resonate the final. With 1400 volts on the plate, I could load my unit to 200 mA plate current; screen current was 30 mA. This corresponds to 280 watts dc input – power output, measured with a Bird thru-line wattmeter, was 178 watts, representing 64% efficiency.

Although I have not experienced any

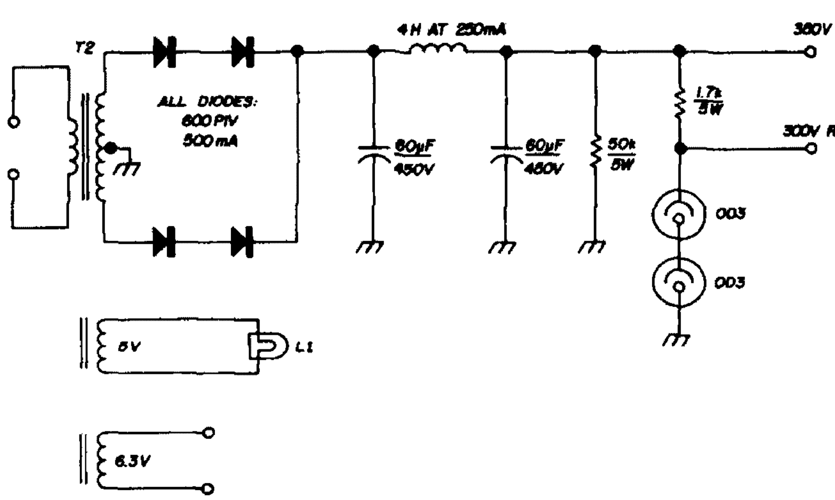


fig. 10. Low-voltage power supply.

instability in my unit, if you should run into this problem, conventional stub neutralization should cure it.

Alignment of the receiving converter is simple and straightforward. Since the mosfets are high-impedance devices, the resonant circuits can be dipped before the transistors are put in their sockets. When the converter is installed in the TR-50, only slight touchup is required to peak all the tuned circuits for maximum output.

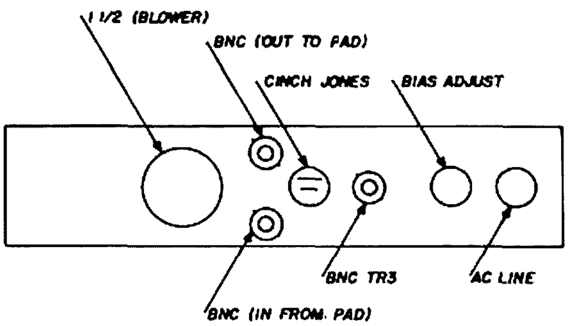


fig. 12. Rear panel layout of the TR-50.

painting

After all the checks and tests were completed and the transverter was fully checked out on the air, the front panel was removed. The steel panel was sanded with fine sand paper using *only* horizontal strokes, followed by a thorough cleaning with a degreasing solvent. *Rust-O-Leum* Dove Gray and Equipment Gray enamel were chosen as the closest match to the Drake TR-3 panel colors. A 3/16-inch wide piece of masking tape on the bare steel divided the panel into vertical halves. Each half was alternately masked while the other half was sprayed. Repetitive light coats of paint are preferred over one heavy coat.

Each color was baked for 30 minutes in the hot-air plenum chamber of a forced hot-air home furnace. After each half of the panel was baked, the 3/16-inch masking strip was carefully removed, leaving a bright steel dividing line that matches the decorative strip on the TR-3.

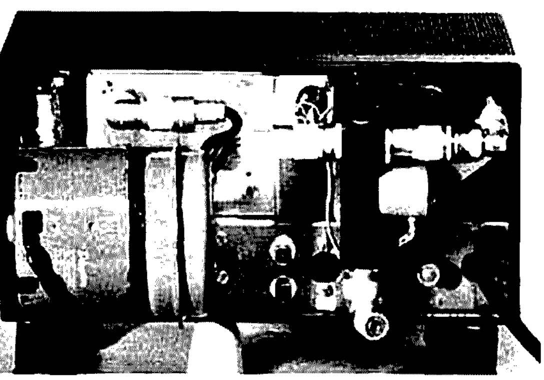
For lettering on the front panel I used *Prestype* transfers — black letters on the light gray upper half, and white lettering on the lower dark-gray half. Three coats of clear acrylic lacquer spray, baked for 30 minutes, protect the lettering.

The elevation legs for the TR-50 were made from 1-inch dowel rod, capped with rubber crutch tips.

conclusion

The original design was recently modified when I acquired a 4X250R from a friend. The 4X150 was replaced, and the bias readjusted for 125 mA idling current. No other modifications were necessary. An immediate increase of 20 watts was noted in output power. This tube, although expensive, is designed specifically for class-AB linear service, and the increase in power and efficiency was well worthwhile.

The pleasure I have derived from designing and building the TR-50 has been multiplied many times by the excellent on-the-air reports I have received. The excellent results have spurred me to begin another project — a TR-144 using K2ISP's printed-circuit two-meter mixer.



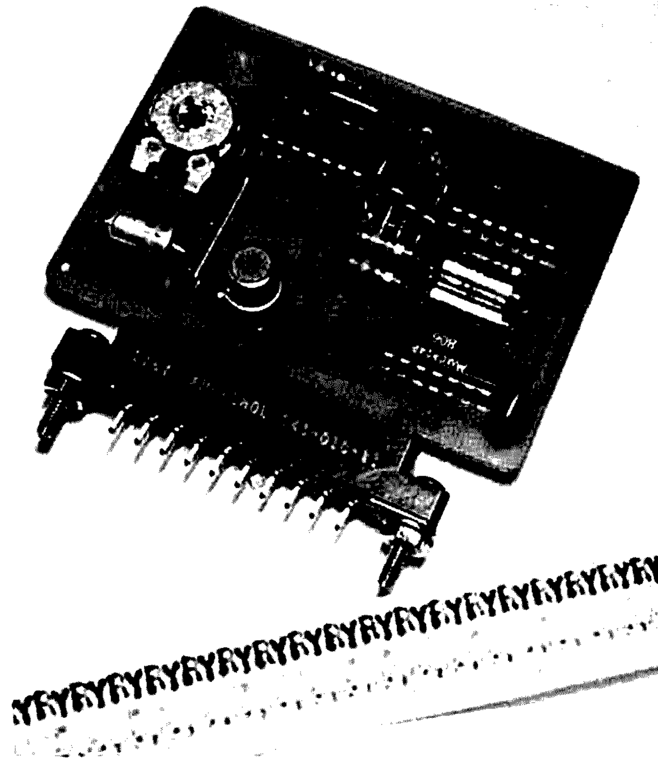
Rear view of TR-50 shows blower and Dow-Key antenna changeover relay.

I would like to express my thanks to Dick McGinn, WA1IMS, for providing the excellent photographs of the TR-50.

references

1. D. W. Bramer, K2ISP, "Heterodyne Transmitting Mixers for Six and Two Meters," *ham radio,* April, 1969, p. 8.
2. D. W. Nelson, WB2EGZ, "Six-Meter Mosfet Converter," *ham radio,* June, 1968, p. 22.

ham radio



RTTY signal generator

This integrated-circuit RTTY reversal generator produces an RYRYRY signal for testing teleprinters

Vic Shinsel, W7ZTC, 2116 Oak Street, LaGrande, Oregon 97850

Teleprinter range adjustments and other performance checks are usually made while the machine is printing the standard RY test signal. *R* signals select even mark and odd space pulses, then *Y* signals reverse everything and select even space and odd mark. The result is a maximum amount of slipping and sliding of adjacent parts within the machine, and conditions which are likely to cause misprint are readily observed.

Bias and distortion measurements are ordinarily taken during reception of *reversal* square-waves or continuous space-bar signals. Reversal signals alternately open and close the machine loop, simulating a succession of mark and space pulses. Space-bar signals perform the same job if a keyboard of known quality is available. Use of the reversal square-wave or keyboard space-bar signal to a questionable RTTY transmission unit promptly turns up any bias distortion – as evidenced by alteration of the

pulse duty cycle. This is the electronic counterpart of the mechanical slipping and sliding check.

Here's a low-cost reversal generator which also sends repetitive, zero-bias, RYRYRY test signals to the printer or transmitter loop. An electronic clutch is included, making the unit a self-contained solid-state transmitter-distributor. A *clutch keyer* circuit may be added later if desired. This enables the generator to send "... RY (pause) RY (pause) RY . . . ," leaving both hands free for stubborn machine adjustments. A multi-purpose RTTY bias and loop test meter will also be described.

In order to be useful in distortion tests, reversal square-wave pulses must be of accurate length, with reasonably fast rise and fall times, and of especially accurate duty cycle. The square-wave works out to be 22.5 Hertz for the 45.5 baud (60 wpm) standard system. Mark and space pulses are of equal duration, so the duty cycle is exactly 50 percent.

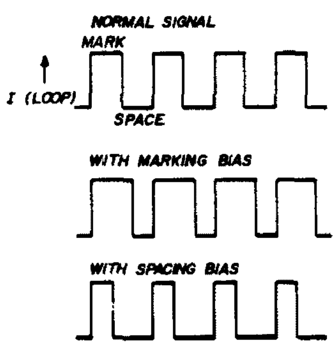


fig. 1. Bias distortion.

theory of operation

Consider a defective demodulator or terminal unit which changes the duty cycle by lengthening the mark pulses and shortening the space pulses. The demodulator is said to cause "marking bias" and it degrades system performance according to the severity of the condition. This kind of distortion (fig. 1) may be one reason why two teleprinters are unable to communicate through a radio system, while both print fine on local loops. The effects of bias distortion may be additive.

Quantitative bias measurements are expressed as percentages. Distortion which lengthens a 22-millisecond mark pulse by 11 milliseconds is described as 50% marking bias.

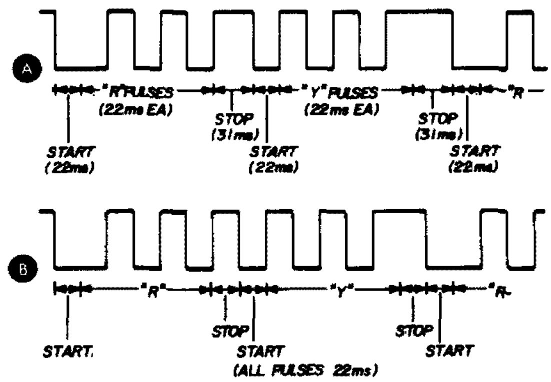


fig. 2. Teletype Corporation RY signal is shown in A; Western Union RY signal is shown in B.

The RY generator takes advantage of a compatibility scheme to develop a signal which prints RY while maintaining the perfect 50% duty cycle of the reversal square-wave. As such, this generator can be a very useful standard for RTTY system tests.

Fig. 2A shows the waveform of a Teletype Corporation model-14 distributor sending RY at the usual 61.3 words per minute. Fig. 2B shows the waveform of a surplus Western Union distributor sending RY at the compatible speed of 65 words per minute. Both machines are usable in the amateur bands, but 65 wpm is right at the upper speed tolerance limit set by FCC.

Note that the only difference between the two signals is the duration of the stop pulses. The Western Union machine sends a shorter (22 millisecond) stop pulse. The two machines operate compatibly because the *selecting* pulses are all the same length, i. e. both units use a 45.5 baud code.

Even with the shorter stop pulse, synchronization is assured. The nominal (420 rpm) receiving shaft speed of the Teletype machine is still faster than the

(390 rpm) transmitting shaft speed of the Western Union machine.

The Western Union RY signal in fig. **2B** may be used as a reversal signal, because each mark has a space counterpart of equal duration. The duty cycle is exactly 50 percent.

circuit

The circuit of fig. **3** generates the RY waveform of fig. **2B.** Gates E and F of the MC824P quadruple-two-input gate are connected as an astable multivibrator. The negative slopes of the multivibrator output trigger flip-flop A (MC890P) to form a perfect square-wave. Gate H is used as a buffer to drive the base of the Motorola MJ421 transistor with the square wave. The MJ421 is rated at 350 volts Vcb and makes an excellent loop driver. Diode D4 is for inductive transient suppression. Resistor R6 (not shown on the photo) was added to decouple transients from the ic logic. The remainder of the circuit is used to invert two out of every fourteen square-wave pulses to form the RY test signal.

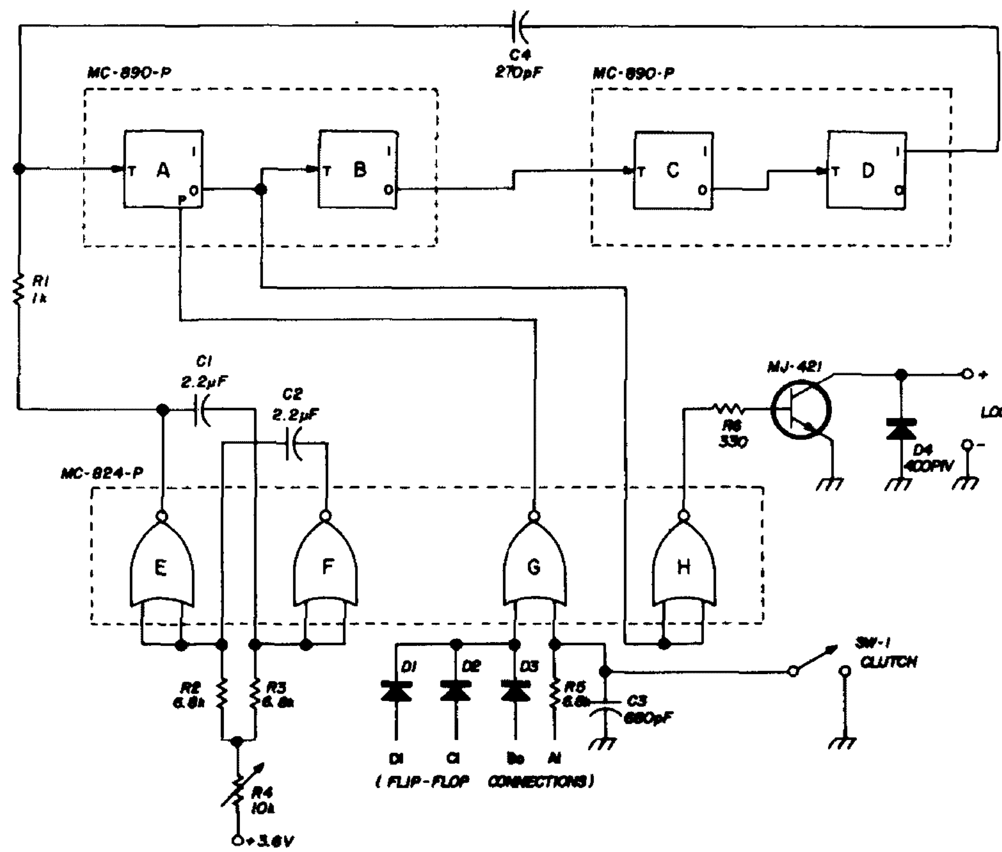


fig. 3 Solid-state RY generator. Diodes D1, D2 and D3 are germanium computer types. Silicon diodes may be used with higher supply voltage (3.5 to 4.5 volts). Resistor R4 is a Mallory MTC1414.

Flip-flops A, B, C and D form a scale-of-sixteen counter. Feedback capacitor C4 inverts one pulse by causing count 2 to be skipped. This occurs during the decay of the positive pulse fed through the capacitor from flip-flop D, which retriggers A and causes an immediate advance to count 3. Gate G inverts the other pulse by presetting count 15 to count zero. In other words, count 15 is also skipped. The truth table of fig. 4 summarizes generator operation.

When S1 is closed, the circuit advances to the mark pulse of count 14. Gate G recognizes this count because all inputs are zero. The output of gate G goes positive and presets flip-flop A to mark. The circuit holds the mark as long as S1 remains closed, because the preset line

will not allow flip-flop A to toggle. This provides the clutch action for the generator.

When S1 is opened, the positive voltage at R5 (from A1) pulls down the output of gate G and removes the preset condition from the first flip-flop. The next trigger from the multivibrator momentarily advances the circuit to count 15. Then the positive voltage disappears from R5, causing the output of gate G to rise. The positive output from G rapidly reactivates the preset line and forces the circuit on to count zero. Gate G has no further influence until count 15 is reached again, (or count 14 if S1 gets closed). Capacitor C3 adds a delay of 0.5 microsecond to ensure that the preset pulse is of sufficient duration to do its job.

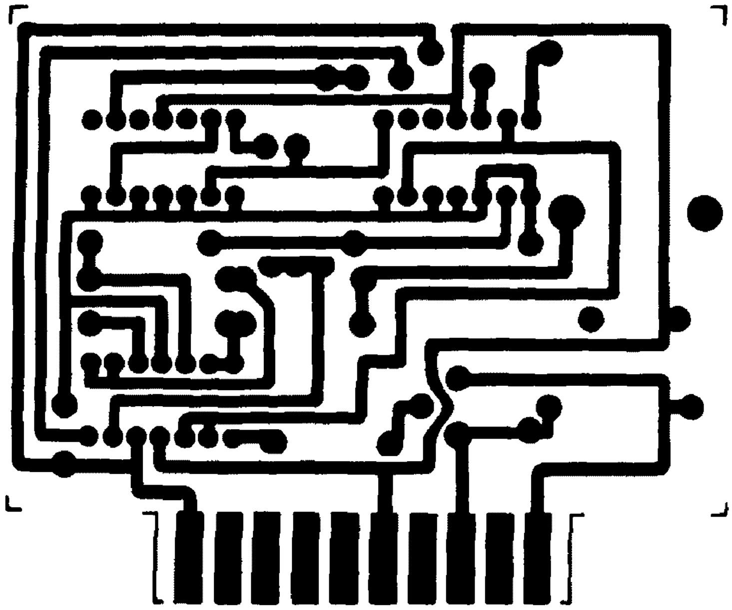


fig. 5. Full-size printed-circuit board layout for the RTTY generator.

fig. 4 Truth table for RTTY generator.

Count	Ao	Bo	Co	Do	Output
0	0	+	+	+	Mark (Stop)
1	+	+	+	+	Space (Start)
2	0	0	0	0	pos. transient — 0.5 µsec.
3	+	0	0	0	Space
4	0	+	0	0	Mark
5	+	+	0	0	Space R
6	0	0	+	0	Mark
7	+	0	+	0	Space
8	0	+	+	0	Mark (Stop)
9	+	+	+	0	Space (Start)
10	0	0	0	+	Mark
11	+	0	0	+	Space
12	0	+	0	+	Mark Y
13	+	+	0	+	Space
14	0	0	+	+	Mark
15	+	0	+	+	neg. transient — 0.3 µsec.
0	0	+	+	+	

The overall effect is that *only* complete RYs are sent. The circuit action is always stopped in the proper state to begin another RY. If S1 is opened, then closed, a single RY will be sent. If S1 is opened, and remains open, a continuous train of RYs is sent. If S1 is eventually closed, the unit finishes its last RY, then stops ready to send another.

construction

Construction of the RY-reversal generator is simplified by the full-size etched-circuit layout of fig. 5. Component locations are given in fig. 6. This is a compact board, and every caution must be taken to avoid solder bridges and excessive heat. It is recommended that all hole locations be dimpled with a sharp scriber, then drilled out to no. 60 holes. Enlarge the pot mounting holes to no. 40 and 50, as required. Brighten the copper with steel wool or SOS pad just prior to soldering.

Surplus flatpack ics worked fine on

the original breadboard, but layout clearances were bothersome. This board uses popular plastic dual-inline RTL ics. The pattern was designed to match an Amphenol 143-010-01 connector. Three to

multivibrator. Use 60-Hz line sweep and set the speed pot for minimal motion of the Lissajous pattern. This will be a slight adjustment! The Lissajous ratio is 22.5 : 60, or 3/8.

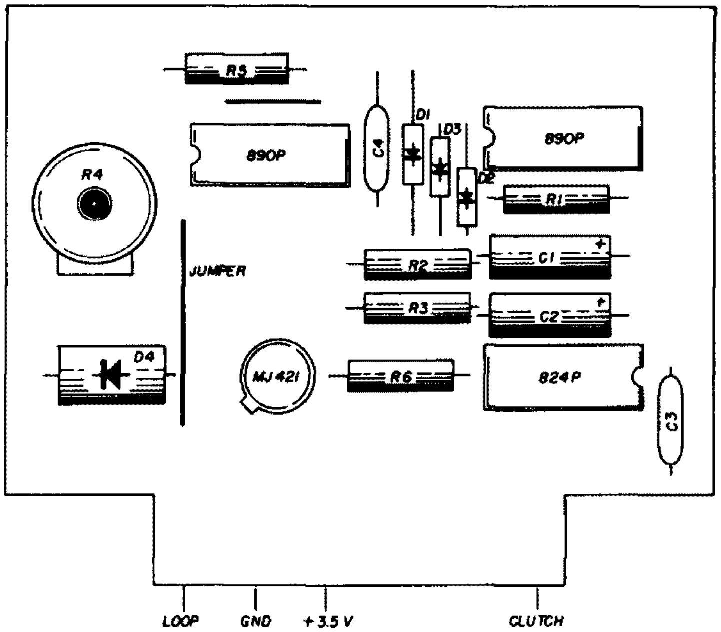


fig. 6. Component placement on the printed-circuit board.

four volts at about 75 milliamperes are required to power the unit; two flashlight cells are fine!

Ham printer loops are capable of some pretty wild transients. Some may develop spikes as high as 800 volts. Avoid ground loops and long leads common to the loop and board supply. A particularly noisy loop may require an 0.1 μF disc-ceramic capacitor from the board +3V to ground right at the Amphenol connector. This was not required on the original unit. Be sure to observe polarity when connecting the generator into a loop, or it will fail to print. No damage should result from a reverse connection.

adjustment

Only one adjustment is necessary. Open S1 and set the speed control (R4) for an approximate 22.5 Hz square-wave at the output of the multivibrator. The printer will start typing RYRY as the proper frequency is approached. Then, for precise adjustment, connect an oscilloscope probe to the output of the

Other speeds may be set on the pot for local machine tests, but it is wise to observe the FCC limits of 55-65 wpm for on-the-air tests.

There is a certain comfort in having nice, clean zero-bias RYs available at the flick of a switch. Automatic carriage-return shows its worth with this unit slamming out continuous RYs.

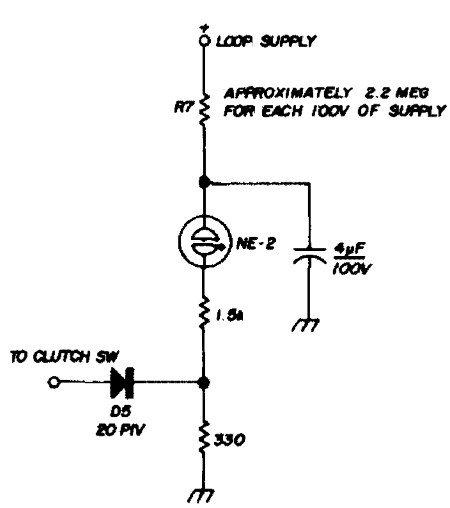


fig. 7. Clutch-keyer circuit.

clutch keyer

The clutch keyer of fig. 7 was an afterthought. Note that the RY generator's clutch-control line is grounded through the diode and 330-ohm resistor. The ground inhibits generator operation. Every few seconds the neon relaxation oscillator fires and momentarily reverse-biases the diode, which effectively removes the ground from the clutch. As a result, one single RY is sent at a time – every two seconds. R7 is about 2.2 meg for each 100 volts of loop supply.

measuring distortion

A common scheme for measuring bias distortion is as follows. First, close the loop (steady mark) and measure the loop current. Then apply reversal signals to the unit under evaluation, and again measure loop current. Since the reversal signal is half mark and half space, the meter will flutter slightly – but its average indication should be exactly half of the closed loop reading. This corresponds to zero bias, and any variation indicates the presence of gremlins. This seems to be a popular method because just about any loop milliammeter is suitable for the test. But there is a pitfall.

The regulation of the loop supply may influence the bias reading. This difficulty popped up in my Mainliner. The mark pulses had a high instantaneous current when first keyed on. Then the current drooped down slightly to the steady-mark value. Fig. 8 shows the loop current waveform through a resistive load. With this condition, the meter errantly indicated a duty cycle change. The triangular "droop" portion of the signal above the steady-mark (60 mA) value appeared as marking bias. In this particular case, adjustment of a slicer or polar relay to reduce the apparent bias would have introduced about 8% distortion.

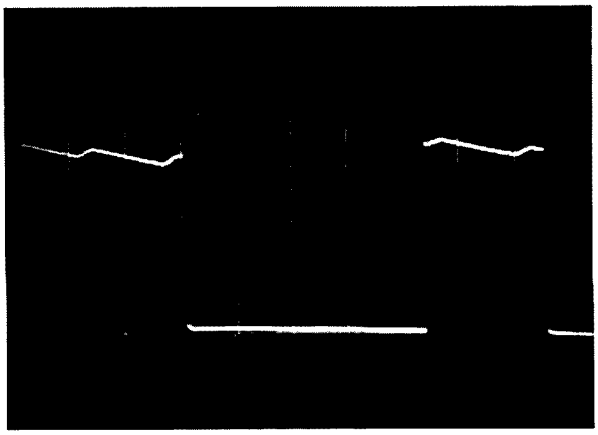


fig. 8. Poor regulation of loop current into resistive load. Droop above 60-mA value looks like marking bias to milliammeter. (Vertical scale, 20 mA per division; horizontal scale, 10 msec per division.)

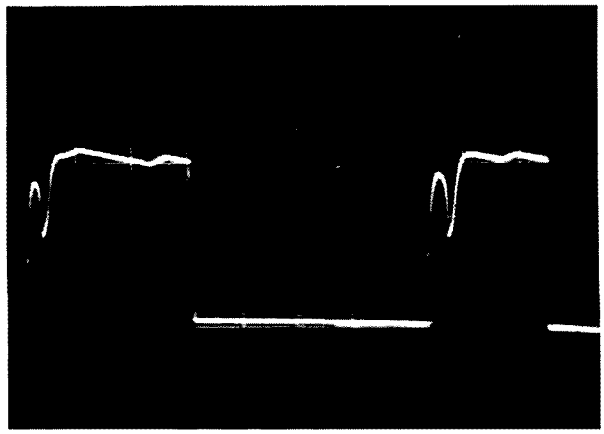
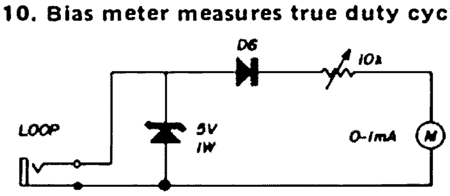


fig. 9. Inductive transients from selector magnet may also fool the bias milliammeter. (Vertical scale, 20 mA per division; horizontal scale, 10 msec per division.)

The situation worsens as inductive loads are switched into the loop. Fig. 9 shows the effect of switching in a model-26 selector magnet.

A better bias measuring scheme is given in fig. 10. The zener diode clips at 5 volts and standardizes the mark pulses. Now the meter shows true duty cycle, and the reading is not appreciably affected by poor regulation, loop transients, loop resistance or supply voltage.

fig. 10. Bias meter measures true duty cycle.



D6 = SILICON COMPUTER DIODE, 20 PIV

The pot is set once and for all at full scale on steady mark. Diode D6 assists in transient suppression.

The meter may be calibrated directly in bias percentage if desired. Zero current corresponds to 100% spacing bias; half scale is zero bias; and full scale is 100% marking bias. Fig. 11 is a suitable scale for a three inch Simpson movement. The scale of fig. 12 was developed for the multi-purpose RTTY meter shown in fig. 13. It measures loop current, loop supply voltage and bias.

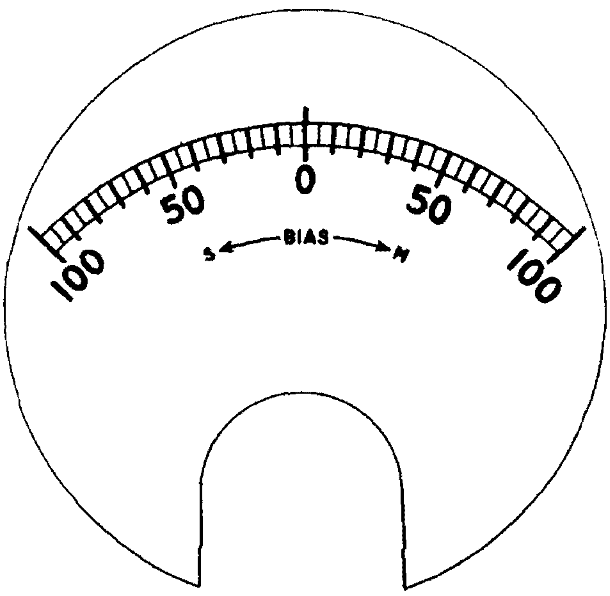


fig. 11. Full-scale bias scale for 3-inch Simpson meter.

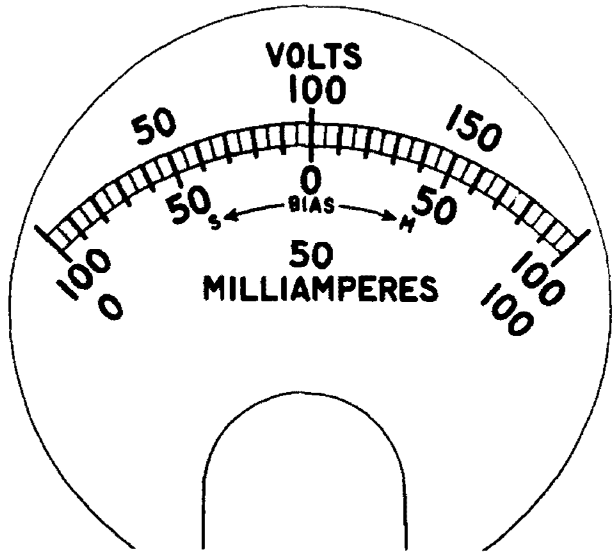
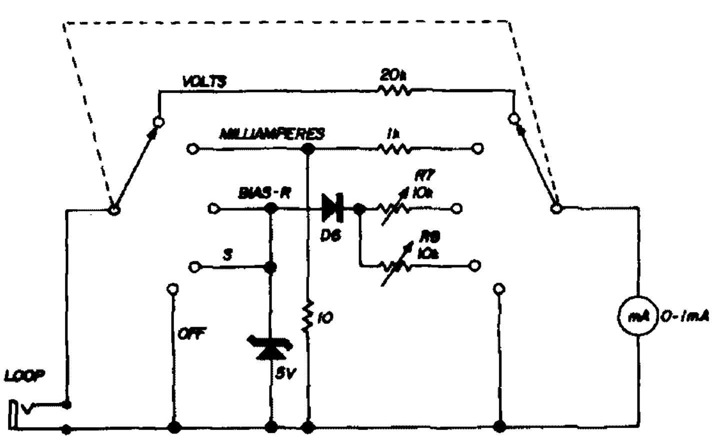


fig. 12. Full-scale RTTY multimeter scale for 3-inch Simpson meter.

A steady stream of keyboard-space signals has a duty cycle of 31% for the standard Teletype equipment, and 35% for the Western Union machine. Potentiometer R8 translates a compromise duty cycle of 33% to center scale on the multimeter for bias-distortion measurements where no reversal generator is available. If the distributer is known to be free of bias, simply send successive keyboard-space signals at full speed. Read the bias at the printing end for an overall check.

For calibration: Set R7 to full scale on steady mark. Then set R8 to 66% of the resistance or R7. As an alternative, R8 may be set to zero bias on a good, synchronous transmitter-distributor. When R7 is in the circuit, the meter reads bias from reversal signals. When R8 is in the circuit, the meter reads bias from keyboard-spaces.

fig. 13. Circuit for RTTY multimeter.



bibliography

1. Irv Hoff, "Operating the Teleprinter," *QST*, February, 1965, p. 29.
2. Charles Teeters, "Teletype Reversal Generator," *CQ*, March, 1958, p. 41.
3. Durward Tucker, "RTTY From A to Z," *CQ*, August, 1964, p. 27.
4. John Conners, "Photographic Printed Circuit Process," *73*, August, 1968, p. 14.
5. Hank Olsen, "Micro-logic for Non-Logical Users," *73*, June, 1967, p. 58.

ham radio

tabulated characteristics

of vacuum-tube and transistor amplifiers

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Although there are many active devices that are used for generating or amplifying af and rf signals, only two, vacuum tubes and transistors, are of great interest to radio amateurs. These two can be split into many sub-categories. However, only the conventional vacuum tube in the common-cathode configuration and the bipolar transistor in the common-emitter configuration will be considered in this article.

Transistors are finding great favor among both manufacturers and amateurs for use in receivers and low-level stages of transmitters. The vacuum-tube amplifier, however, still ranks as the best choice for moderate and high-power stages in radio transmitters.

Intelligent use of either device requires that the user have a clear concept of the device's capabilities, especially the limitations of these capabilities. Then, too, the FCC likes to ask rather searching questions concerning both vacuum tubes and transistors in its amateur license examinations. With the upsurge of interest in qualifying for higher-grade licenses, these active devices are having a new and inquiring look taken at their characteristics and capabilities.

Many books and magazine articles have been published on the characteristics and applications of vacuum tubes; almost as many have been *written on* transistors. Most of these are very good for the study of a particular aspect of the subject. A tabulation of the relative characteristics of the three major classes (A, B, and C) of amplifier operation has not had the emphasis that interest demands. This aspect must be understood clearly if an amateur is to make knowledgeable use of such amplifiers, whether in equipment of his own design or in the correct adjustment of manufactured equipment.

This article is written to provide a simple but complete comparison of the operating characteristics of class-A, class-B, and class-C amplifiers, listed under one heading for vacuum tube amplifiers and under another for bipolar transistors. Class AB1 and class AB2 operation may be interpolated between class A and class B.

This tabulation makes no pretense of being an in-depth exposition of the behavior of vacuum-tube amplifiers, and most certainly, not of transistor amplifiers. The facts, however, concisely state the more important factors that determine the use to which we put such amplifiers, and how we operate them in those applications.

table 1. Characteristics of vacuum-tube amplifiers.

	Class-A Operation
Distinguishing property	Ability to amplify signals without distortion.
Desired effect	To have the waveform of the output signal be identical, except for amplitude, to the waveform of the input signal.
Method	Quiescent grid-bias voltage is set to establish the operating point midway on the straight portion of the E_cI_b characteristic curve. Grid signal voltage is limited so it never goes beyond the straight portion of the E_cI_b characteristic curve in either the negative- or positive-going extremes of signal voltage. For triode vacuum tubes, for maximum undistorted output, the plate load resistance is set at twice the dynamic plate resistance of the tube. Plate voltage is limited to a value that will ensure the tube's rated plate dissipation will not be exceeded.
Distinguishing characteristics	Voltage gain is very high Power gain is very high Grid drive power requirement is extremely low Plate current, as measured with a dc meter with signal present, remains constant Plate current flows 360° during an input signal cycle Input impedance is very high Output impedance is not critical; however, distortion increases more for too-low than for too-high mismatch No grid current flows Grid-bias voltage source does not have to have low impedance or good voltage regulation Plate voltage source does not need to have good voltage regulation The stage may be used either single-ended or push-pull for either af or rf signals Stage can not be used as a frequency multiplier Oscillators can be forced, by limitation of grid swing, to operate in this class Efficiency is approximately 20%

Class-B Operation

Distinguishing property Ability to amplify amplitude-modulated signals without distortion of the modulation characteristics.

Desired effect To have the output signal power vary as the square of the input signal voltage.

Method Quiescent grid-bias voltage is set to establish the operating point just short of the cut-off point on the E_cI_b characteristic curve.

Grid signal voltage may be great enough to drive the grid into the positive region. The less it is driven into the positive region, the lower will be the distortion of the output signal.

Plate load resistance is adjusted to the value required to attain the desired effect. Normally, it is adjusted for very heavy plate current loading.

Plate voltage is limited to a value that will ensure the tube's rated plate dissipation not being exceeded even when the plate current is set high to attain the desired effect.

Vacuum tube is selected to have ample cathode emission to meet heavy peak current demands.

Distinguishing characteristics Voltage gain is moderate

Power gain is moderate

Plate current, as measured with a dc meter with signal present, may vary in relation with signal voltage amplitude. This is the condition when amplifying af or ssb rf signals. When amplifying a-m (double sideband with carrier), fm, or key-down cw signals, the plate current remains constant

Plate current flows for approximately 180° during an input signal cycle

Input impedance varies from high, when the grid is not excited into the positive region, to low, when the grid is driven into conduction

Output load impedance is quite critical, as its optimum adjustment is an important factor in achieving the desired effects

Grid current may flow; when it flows, it varies in the same manner as does the plate current

Grid-bias voltage source needs low internal impedance and good voltage regulation

Plate voltage source needs good voltage regulation

Stage makes a moderately good frequency multiplier

Stage must have push-pull configuration for af use but may be either single-ended or push-pull for rf

Oscillators can be forced to operate in this class

Efficiency is approximately 30% to 33% for steady-state rf (double sideband with carrier, fm, or key-down cw) and may go as high as 66% for af or ssb rf

Class-C Operation

Distinguishing property	Ability to respond linearly to amplitude modulation of the plate circuit
Desired effect	To have the plate current vary directly with the plate voltage. (The plate current, for example, will double when the plate voltage is doubled)
Method	Quiescent grid voltage is set to establish the operating point at approximately two-and-one-half times the negative potential required to cut off plate current flow. A portion of this voltage may come from a fixed source; part, however, must come from the IR drop across a grid resistor
	Grid signal voltage must be great enough to drive the grid far into the positive region, preferably into saturation
	Plate load resistance is not critical; it usually is set by complying with the requirement for moderate plate loading
	The vacuum tube is selected to have ample cathode emission to meet heavy peak current demands
	Plate voltage is limited to a value that will hold the plate dissipation to well within its rated capabilities, preferably to not over 60% of the rate capability; also, the plate voltage must not exceed one-half the flash-over voltage rating for the tube
Distinguishing characteristics	Voltage gain is very low
	Power gain is low, averaging about 10
	Grid drive power requirement is high
	Plate current, as measured with a dc meter under modulation conditions, remains constant
	Plate current flows between 60° and 90°, depending on bias and drive, during an input signal cycle
	Input impedance is low
	Output impedance is not critical; generally it is established by the requirement that the plate current should not exceed about 60% of the tube's rated capability
	Heavy grid current flows
	Grid-bias voltage source does not need good voltage regulation
	Plate voltage source needs good voltage regulation
	Efficiency is approximately 65% to 75%
	The stage may be used in either single-ended or push-pull configuration
	The stage makes an excellent frequency multiplier
	Oscillators normally operate in this class

table 2. Characteristics of bipolar transistor amplifiers.

Class-A Operation

Distinguishing property	Ability to amplify signals without distortion
Desired effect	To have the waveform of the output signal to be identical, except for amplitude, to the waveform of the input signal.
Method	Quiescent base current is set to establish the operating point midway between the cut-off and saturation points on the ac loadline
	Base signal voltage is limited to produce equal variations of collector current for both positive- and negative-going excursions of base-emitter signal voltage
	Load resistance is made equal to the value of the collector-emitter voltage divided by the collector current
	Collector-emitter voltage is limited to a value that will not permit the transistor's dissipation to exceed the rated power under any condition of signal voltage
Distinguishing characteristics	Voltage gain is very high
	Power gain is high
	Base-emitter drive power is moderate
	Collector current, as measured with a dc meter with signal present, remains constant
	Collector current flows 360° during an input signal cycle
	Input impedance is quite low
	Output impedance is not critical
	Base-emitter current flows 360° during an input signal cycle
	Collector voltage source does not need good voltage regulation
	The stage may be used either single-ended or push-pull for either af or rf signals
	The stage cannot be used as a frequency multiplier
	Oscillators can be forced, by limitation of base-emitter voltage swing, to operate in this class
	Efficiency is low, running from 20% to 30%

Class-B Operation

Distinguishing property	Ability to amplify amplitude-modulated signals without distortion of the modulation characteristics
Desired effect	To have the output power vary as the square of the input signal voltage
Method	Quiescent base-emitter current is set to establish the operating point just short of collector current cutoff For rf linear-amplifier service, a transistor is selected to have: 1. A flat beta curve 2. Multiple emitters, each with an integral bias resistor 3. High second-breakdown characteristic 4. Power dissipation at least twice the power at which it will be operated For rf linear-amplifier service, the peak collector current swing is limited to only 30% of the rated maximum current Collector load resistance is adjusted to the value required to attain linearity; normally, it is adjusted for very heavy collector current loading
Distinguishing characteristics	Voltage gain is moderate Power gain is moderate Collector current, as measured with a dc meter with signal present, may vary with signal voltage amplitude Collector current flows for approximately 180° during the input signal cycle Input impedance is low Output impedance is quite critical, as its optimum adjustment is an important factor in achieving the desired results Base-emitter current will flow during slightly over 180° of the input signal cycle Base-bias source must have low internal resistance Collector voltage source must have good voltage regulation The stage makes a good frequency multiplier The stage must have push-pull configuration for af use but may be either push-pull or single-ended for rf use Oscillators can be forced to operate in this class Efficiency varies from 30% to 66%

Class-C Operation

Distinguishing property	Ability to respond linearly to amplitude modulation of the collector circuit
Desired effect	To have the collector current vary directly with the collector voltage. (The collector current, for example, will double when the collector voltage is doubled)
Method	Quiescent base-emitter current is zero; the base-emitter voltage is biased so that a small signal voltage is required to cause base-emitter (and collector-emitter) current to flow
	Input signal voltage must be great enough to drive the transistor to collector saturation. Because of the decreasing efficiency of a transistor under high excitation, the input signal also must be amplitude modulated. This serves two purposes: it supplies the extra peak excitation to keep the output power up during upward modulation peaks, and decreases the excitation to lessen feed-through during downward modulation swings
	Collector load resistance is not critical
	Collector voltage must be limited to not over one-half, and preferably not over one-fourth, the rated maximum voltage
Distinguishing characteristics	Voltage gain is very low
	Power gain is low, from 3 to 10 dB
	Base-emitter drive power is quite high
	Collector current, as measured with a dc meter under voice modulation conditions, remains constant
	Collector current flows between 90° and 160°, depending upon base bias and drive, during an input signal cycle
	Input impedance is very, very low
	Base-emitter current is high under signal conditions
	Base-bias source does not need good regulation
	Collector voltage source requires good regulation
	Efficiency is approximately 65% to 75%
	The stage is useful only in rf service and may be used either single-ended or push-pull
	The stage makes an excellent frequency multiplier
	Oscillators normally operate in this class

ham radio

plain talk about repeater problems

Intermodulation and desensitization are the two most serious problems faced by repeater users — here is a discussion of some solutions

Vern Epp, VE7ABK, 203 View Street, Nelson, British Columbia, Canada

There are presently over 300 vhf amateur repeaters in use in the USA and Canada. Although a number of articles have been written on the subject, very little has been published about some of the problems. Basically, a repeater receives a signal on one frequency and transmits it on another. A carrier-operated relay is normally used to turn on the transmitter with an incoming signal. One of the biggest repeater problems is desensitization intermodulation and transmitter noise in the receiver. With proper design this problem can be alleviated.

The transmitter interferes with the receiver in two ways. First, the high signal power overdrives the grids of the receiver rf mixer or i-f tubes and causes grid current to flow. This grid current produces bias voltages which reduce receiver gain; this is desensitization. Secondly, the output of the transmitter consists of a carrier, sidebands and noise. Since the noise spreads to either side of the carrier, a portion of this noise spectrum falls on the receiver frequency, causing further receiving problems.

For successful operation, the following items must be kept in mind when designing a repeater: frequency spacing, antenna separation and transmitter power.

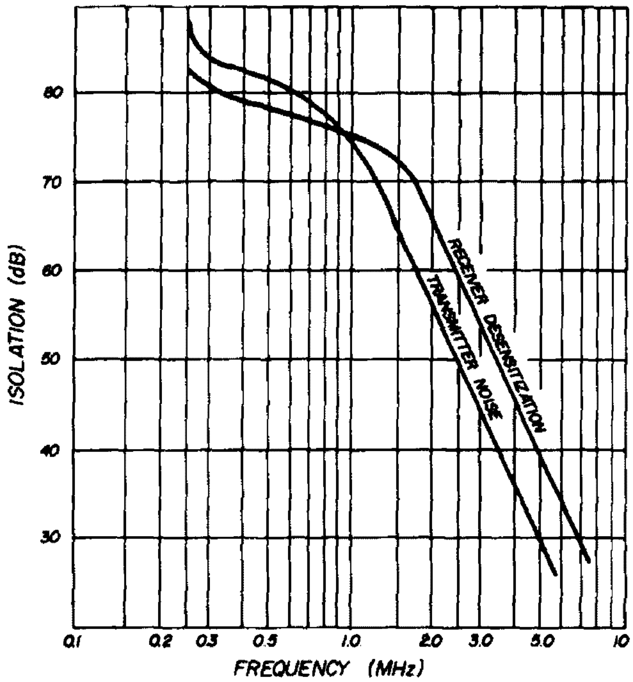

fig. 1. Isolation between transmitter and receiver to protect the receiver from transmitter noise and desensitization.

ceiver desensitization. With very little separation, the effect of transmitter noise is predominant. At greater frequency separations, the effect of receiver desensitization becomes greater (fig. 1).

It is generally agreed that the minimum frequency spacing should be 550 kHz if equipment is physically in one location and two antennas are used. The most common amateur repeater receives on 146.34 MHz and transmits on 146.94 MHz, a separation of 600 kHz. An even larger frequency separation is desirable to reduce the problems discussed.

antenna separation

Antenna separation is a very important factor. Most repeaters use vertical polarization since the mobile stations use vertical antennas. One solution to this problem is shown in fig. 2A where the transmit and receive antennas are physically remote from each other. An interconnecting link on 220 or 450 MHz connects the units together. This system is expensive but antenna isolation is nearly ideal.

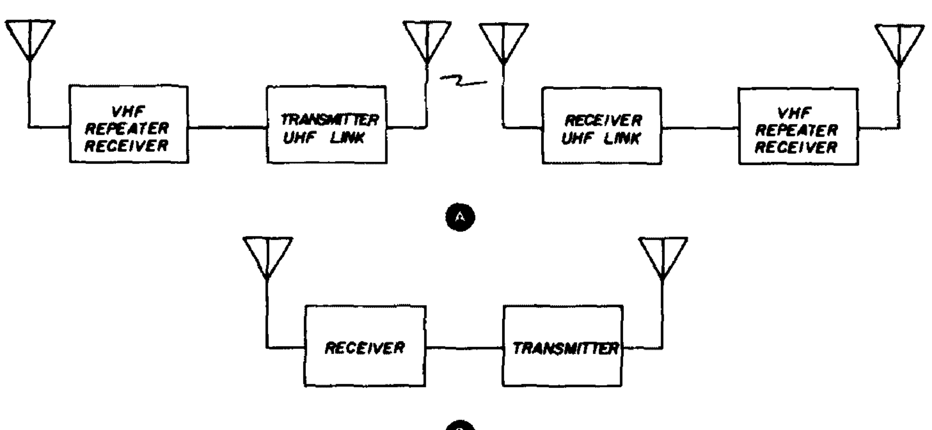

fig. 2. Use of separate antennas for transmitter and receiver.

frequency spacing

The closer the frequency of the transmitter is to the receiver, the more severe the effects of transmitter noise and re-

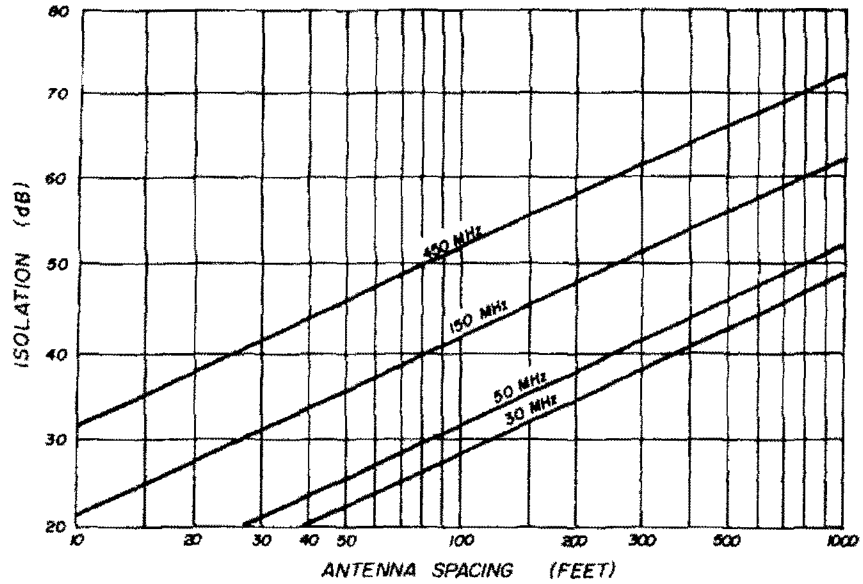

fig. 3. Attenuation provided by horizontal separation of dipole antennas.

The more common, less expensive method uses one location (**fig. 2B**). The antennas are separated vertically or horizontally from each other. **Figs. 3** and **4** show relative attenuation using vertical or horizontal separation. Vertical separation provides better isolation and is recommended for repeater operation. The top antenna is normally used for receiving while the lower one is used for transmitting.

transmit power

Interference problems increase as the power is increased. Most repeaters operate in the 25 to 50 watt range since mobiles use about the same power outputs. There is no point in running high power unless the mobile and base stations are also high powered. Obviously, a repeater gains little with a transmit range that is much greater than its received range.

cavity filters

Cavity filters are excellent devices for reducing interference problems. Loops are used to insert and extract energy from the cavity. The degree of coupling can be controlled by rotating the loops; this

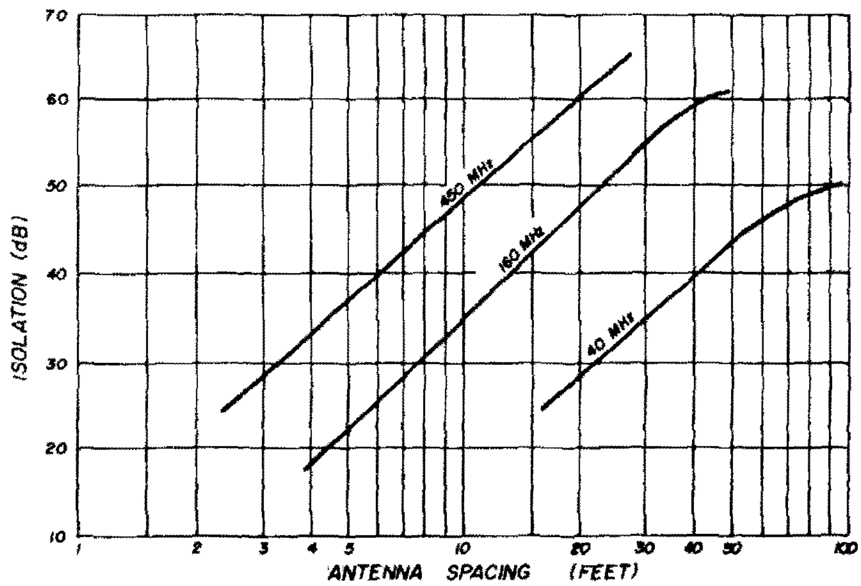

fig. 4. Attenuation provided by vertical separation of dipole antennas.

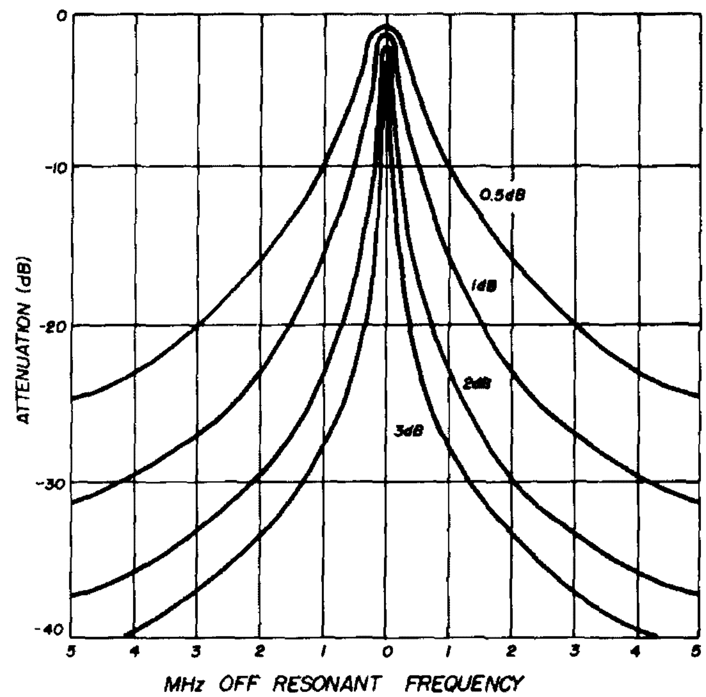

fig. 5. Selectivity curve for Decibel Products DB-4001 single cavity.

determines cavity insertion loss and selectivity. The more selective the cavity, the higher the insertion loss.

When a bandpass cavity is installed in the output of a transmitter it will pass the carrier and sidebands and attenuate nearly all other frequencies. The amount of attenuation depends on the frequency spacing between the undesired frequency

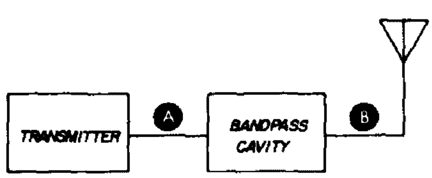

fig. 6. Using a bandpass cavity with a transmitter. Cable length A should be an odd multiple of a quarter wavelength.

and the carrier frequency. A typical bandpass cavity frequency vs attenuation curve is shown in fig. 5, and a typical application is shown in fig. 6.

The cavity presents an impedance of approximately 50 ohms at resonance but the off-resonance impedance of the cavity loop is very low. This makes coaxial cable length (**A** in fig. 6) very important. If the *electrical* length between the cavity loop and the transmitter output circuit is any multiple of a half wavelength, the cavity, if not tuned to resonance, will present a short circuit to the transmitter.

To remedy this, cable length should be an odd multiple of a quarter-wavelength. The transmitter will then see a high impedance when the cavity is off resonance, and danger of transmitter damage is much smaller. Cable length **B** in fig. 6 is not critical.

bandpass cavities in receivers

Fig. 7 shows a cavity installed between the antenna and receiver. Coaxial cable length is not critical. The frontend selectivity of the receiver is improved by rejection of off-frequency signals that would otherwise desensitize the receiver or cause intermodulation interference. If greater selectivity is required, two or more cavities can be connected in *series*.

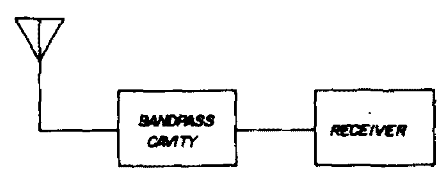

fig. 7. Using a bandpass cavity with a receiver.

checking for interference

1. Switch on the transmitter while receiving a weak signal and watch the limiter current. If it decreases, the receiver is being desensitized. Try additional attenuation in the receiver by adding a cavity.

2. With no signal input, switch on the transmitter. If limiter current increases, transmitter noise is being received. Try additional attenuation in transmitter by adding a cavity.

other items to consider

The coax cables between transmit and receive antennas should never be run together physically. The shielding of RG-8/U, for example, is only about 80%

efficient, and some coupling will result. Also, the transmitter and receiver should be physically separated in the mounting bay. Both units should be well grounded and may require shielding if a problem exists locally and not by the antennas.

Another good method of reducing receiver desensitization was described by W5KPZ.[1] He places a crystal across the antenna coil of the receiver. The crystal is an overtone type that is resonant at the frequency of the interfering signal (fig. 8).

duplexers

A duplexer is used to permit operation of a transmitter and receiver on one antenna. It uses a number of resonant cavities to provide sufficient isolation between units. At close frequency spacings duplexer adjustment is very critical to obtain sufficient isolation. Duplexers for close spacing (0.5 to 1 MHz) commercially cost anywhere from $500 to $1000.

rf interference

Rf interference is occurring more frequently since the frequency spectrum is becoming more and more crowded. Many mountain-top sites have many repeaters located near each other and using the same power levels. As a result many interference problems exist.

Rf interference can be defined as, "rf power which interferes with the reception of the desired signal thus causing reduced intelligibility." Receiver selectivity is of the utmost importance since an off-frequency signal can enter the front end of a receiver even though it is rejected later by i-f selectivity. A number of problems can occur, including desensitization, intermodulation and receiver spurious response. Most commercial manufacturers use high-Q rf filters to obtain better rf selectivity.

intermodulation

Intermodulation products (IM) are generated when two or more frequencies are mixed. They can occur in the transmitter, usually in the power amplifier stage, or in the receiver, usually at the first converter. IM products can be reduced by using bandpass or band-reject cavity filters that are tuned to the interfering signal.

table 1. Calculating intermodulation products.

Order	Formulas for Calculating Possible Intermodulation Products
2nd	$A \pm B$
3rd	$A \pm 2B$, $2A \pm B$
4th	$A \pm 3B$, $2A \pm 2B$, $3A \pm B$
5th	$A \pm 4B$, $2A \pm 3B$, $3A \pm 2B$, $4A \pm B$

Table 1 shows the new frequencies which can be produced by intermodulation. The third- and fifth-order products are the ones most likely to cause trouble. And, when a number of transmitters are located in the same general location, the IM problem becomes very difficult and complex.

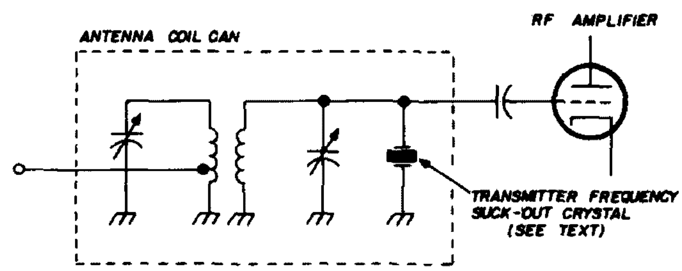

fig. 8. Installation of a suck-out crystal to reduce receiver interference. Crystal is same frequency as that of the interfering signal.

spurious responses

Spurious receiver responses fall into three general categories, i-f response, local-oscillator spurious response and image response.

1. I-f response results when interference at the i-f frequency enters the receiver.

2. Local oscillator spurious response occurs when an off-frequency signal enters the receiver and mixes with a harmonic of the local oscillator to produce an i-f frequency.

3. Image response; the image is twice the i-f frequency plus or minus the desired signal (depending on whether the local oscillator is above or below the incoming signal).

spurious radiation

Transmitters can cause interference to nearby receivers even if it is within the FCC or DOT requirements of 60 or 70 dB down. Interference can come from several sources, including the spurious of a transmitter crystal oscillator, second and third harmonics of the operating frequency, or from the 150-MHz energy in the tripler stage of a 450-MHz transmitter.

intermodulation tests

transmitters If a receiver is interfered with when two transmitters are on the air, place a pad between one of the transmitters and the antenna. If the interference level decreases by twice the dB value of the pad, the undesirable frequency mix is taking place in the transmitter with the pad. This is because the signal from the second transmitter is attenuated on entering the first transmitter, and the intermodulation products are attenuated on the way out. If the intermodulation is taking place in the transmitter without the pad, the interfering signal will be reduced only by the dB value of the attenuator.

receivers If intermodulation is suspected in a receiver, install an attenuation pad in the receiver antenna circuit. For each dB of attenuation in the pad, 3rd order products will be reduced by 3 dB, and fifth-order products by 5 dB. If the interfering signal is generated outside the receiver it will be reduced by the value of the pad.

references

1. D. Baxter, W5KPZ, and J. Douglas, W5OBV, "Two-Meter Repeater," *73*, September, 1966, p. 58.
2. R. C. Trott, "Considerations in the Use of Cavities and Duplexers," Decibel Products, Inc., Dallas, Texas.
3. General Electric Datafile bulletins 10002-2, 10007-2 and 10007-4.

ham radio

the cordover audio-oscillator module

Introducing a completely assembled circuit board that has many amateur applications — for less than a dollar!

J. A. Houser, WB2GQY, 23 Washington Street, Rensselaer, New York 12144

The circuit shown in fig. 1 is undoubtedly one of the best bargains available when one considers the many uses to which it can be adapted.

Known as the Cordover module,* the circuit is designed as an audio oscillator. A pnp low-power audio driver is coupled to an npn 5-watt output stage. The only other parts of the complete module are a 33k resistor and a .02 μF capacitor. As no heat sink is provided, full power can't be obtained without ruining the output transistor. An advantage (or disadvantage, depending on how you look at it) is that the module is designed to work directly into the voice coil of any speaker from 3 to 16 ohms without a matching transformer. Only 3 volts (two size D cells) are needed for efficient operation. Terminals are provided for an on/off switch or key. The circuit will switch with a bug set at its fastest speed.

Audio output is sufficient for most applications; but if greater output is desired, 6 volts may be used at considerably greater battery drain and consequent shorter battery life. I measured the current at 3 volts, and it varies between 50-60 mA over the whole audio range of oscillation. Dc input measures 0.18 watt.

*Available from Allied Radio Shack, bubble-packed with instructions, for 98¢. Listed in their 1971 catalog as no. 20-1155.

The efficiency seems to be close to 50% with just under 100 mW output available.

changing audio frequency

Because many beginning hams or experimenters may not have an audio generator to calibrate the module, and may desire to use it at a frequency other than that at which it's set, I made up table 1 so anyone may choose the desired frequency and substitute the designated capacitor for the one supplied. The tone will be within a few Hz of that specified in the table, allowing for variation in transistors and components. Although specifications mention 1000 Hz (as supplied with the .02-μF capacitor installed), I measured 960 Hz on all modules tested, which seems close enough.

ssb testing

I use these modules for tone-testing an ssb transmitter. Two are used for a 2-tone test at full modulation. (An article is in preparation on this use of the module.) Although the observed audio waveform is closer to a square wave than a sine wave, no difficulty was encountered in evaluating ssb transmitter performance.

other uses

The module can be used as part of an off-the-air monitor;[1] as the subcarrier source for a voice scrambler;[2] and, with an integrator circuit, as a pulse timing source.

spot-frequency applications

Table 2 shows the effect of changing the value of the resistor supplied with the module. Various values of resistance were connected in parallel with the 33k unit. The data in the table was obtained with a .25 μF capacitor in the circuit, which produced the basic (nominal) frequency of 250 Hz. This experiment was made to show how the basic module frequency can be changed by a few Hz to hit a spot frequency between those afforded by various capacitor changes. Five-percent-

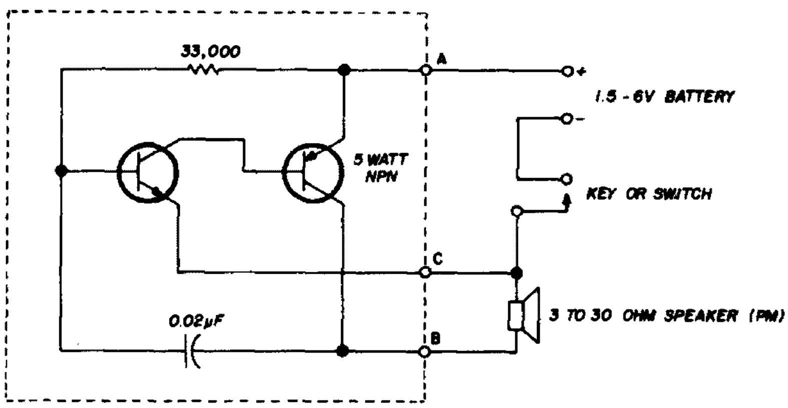

fig. 1. Schematic of the bargain audio oscillator module. Unit comes completely assembled and wired, ready for numerous amateur applications.

table 1. Module frequency as a function of capacitor value. Capacitors shown were substituted for the .02 μF unit supplied with the module. Measurements were taken with 3 Vdc at 50-60 mA.

Capacitor Value (μF)	Measured Frequency (Hz)
40 (electrolytic)	.5
10 (electrolytic)	10
.5	160-170*
.25	280*
.1	400-450*
.05	690
.02	890-960
.01	1440
.005	2800-2900 (harmonic 5700)**
.0025	4400 (harmonic 14,000)**
.001	7000***

* varied with capacitor tolerance.
** harmonic as strong as fundamental.
***out of hearing range but strong.

table 2. Module frequency change as a function of resistor value. Value shown in first column was connected in parallel with the 33k unit supplied.

Resistance (kilohms)	Observed Frequency Increase (Hz)	Measured Frequency (Hz)
1000	5	255
560	10	260
330	15	265
180	25	275
100	50	300
56	105	355
33	175	425

tolerance resistors were used, so a deviation may be found in the numbers given if you use resistors with different tolerances. The frequency can deviate from 1 Hz with a 1 megohm resistor to perhaps 5 Hz with 33k.

harmonic output

Unless the module is fully loaded with a speaker, harmonics and subharmonics can develop at the higher-frequency ranges. This condition can be detected by an increase in current from the nominal (50-60 mA) to 90-100 mA. I ran into this problem when using matching transformers as loads when the transformers were not sufficiently loaded to reflect a low (3-30 ohm) impedance to the transistor output.

fm applications

The information in the paragraphs above suggests the possibility of using these modules for tone-frequency-controlled keyers to actuate two-way frequency-shift mechanisms in certain fm transmitters. Another possibility is their use as function (tone) controls to actuate fm repeater transmitters.

In tests, the Cordover module frequencies were as stable as those obtained in touch-tone telephones. With a little juggling of components, exact zero-beats to these frequencies could be obtained.

No frequency shift was noticeable in any of the ranges when fresh batteries were used. As a matter of fact, frequencies were so stable it occurred to me that eleven of these little modules might be used to construct an inexpensive electronic organ. If more than 1 octave were desired, the modules could be followed with one or more "times two" ICs. The keyboard would probably cost more than the electronics.

Just one of these modules furnishes an excellent audio source for testing mono or stereo amplifiers, from the pickup right through the speakers; PA amplifiers ditto – from microphone or phono input to speakers. An inexpensive 6-ohm to high-impedance miniature transformer comes in handy in such applications.

For 98¢, the Cordover audio oscillator module is hard to beat.

references

1. J. A. Houser, WB2GQY, "A CW Monitor," *73*, March, 1970, page 76.

2. J. Pina, "A Subcarrier Source for a Voice Scrambler," *Popular Electronics*, March, 1970, page 42.

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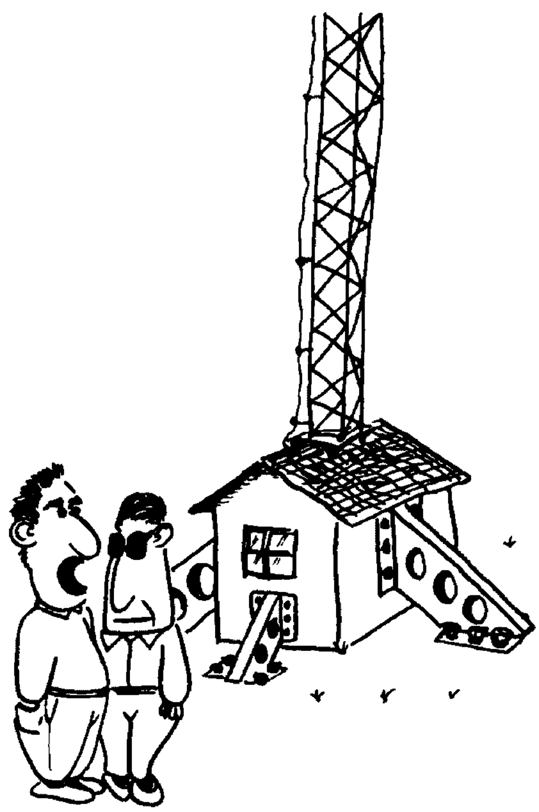


Yeah — the QTH used to rotate about 20° until I put in the shack clamps!

rating tubes

for linear amplifier service

Peak envelope power and intermodulation distortion are important parameters when selecting tubes for linear amplifiers — here's what they mean and how they are measured

Robert I. Sutherland, W6UOV, William I. Orr, W6SAI, Eimac division of Varian

The power-handling capability of a given tube in single-sideband service depends upon the nature of the signal being transmitted and the tube's power dissipating capability. The method of establishing single-sideband service ratings should be such that relatively simple test equipment can be used to determine whether or not a tube is operating within its maximum ratings.

It is impractical to establish a rating based on voice-signal modulation because of the irregular waveforms and the varying ratios of peak-to-average signal power found in different voices. The most convenient rating method, and probably most practical, uses a single-tone audio signal to modulate the ssb transmitter. By using this test signal at its full modulation capability, the amplifier will operate under steady, maximum-signal conditions which are easily duplicated and observed.

When a single sine-wave tone modulates a single-sideband transmitter the rf output appears as a steady, unmodulated signal on an oscilloscope (see fig. **1A**). This is because the output is a continuous signal having a frequency removed from that of the carrier by the modulating frequency, as shown in fig. **1B**.

two-tone tests

Consequently, the operation of a *linear amplifier under single-tone* modulation is comparable to that of a cw transmitter under key-down conditions. As such, the performance of the power-amplifier stage at maximum signal (or peak) conditions can be determined from meter readings. However, this simple test lacks information on the linearity of the

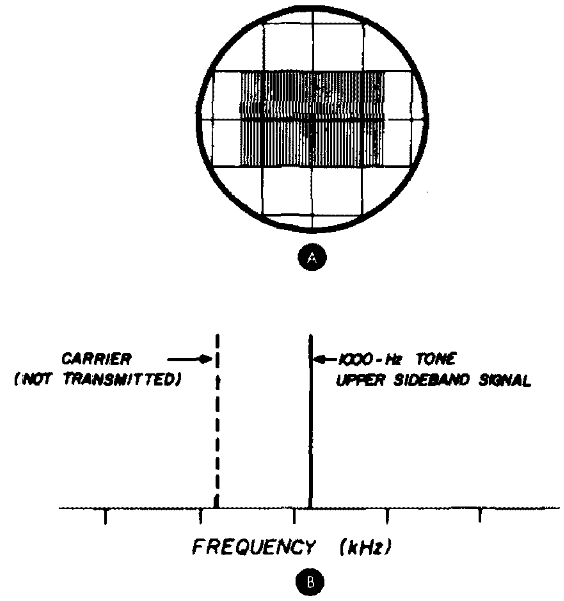


fig. 1. Rf output of ssb transmitter with single-tone modulation. Oscilloscope pattern is shown in A; spectrum is shown in B.

stage. To study linearity by observing amplifier output, some means must be provided to vary the output signal level from zero to maximum with a regular pattern that can be easily interpreted. A simple means is to use two equal-amplitude audio tones to modulate the ssb transmitter. This is termed a *two-tone* test. With this procedure the transmitter emits two steady signals separated by the frequency difference of the audio tones (fig. 2).

In some ssb generators, the two-tone signal is obtained by impressing a single tone at the audio input and injecting the carrier (by unbalancing the balanced modulator) to provide the second equal amplitude rf signal (fig. 3). The resultant beat between the two rf signals produces a scope pattern which has the appearance of a carrier 100 per cent amplitude-modulated by a series of half sine waves as shown in fig. 4.

When using the two-tone technique to measure the distortion of a linear rf amplifier it is sometimes more expedient to use two rf signal sources (separated in frequency by the desired number of cycles) and to combine them in a manner which will minimize the interaction between them. The two rf signals represent the two equivalent sideband frequencies generated by the two-audio-tone system and produce exactly the same scope pattern.

A linear amplifier is usually rated at peak envelope input or output power level. *Peak envelope power* (PEP) is the root-mean-square (rms) power generated at the peak of the modulation envelope. With two-tone or single-tone test signals the approximate relationships between single-and two-tone meter readings, peak envelope power and average power (class B or AB operation) can be determined from the formulas shown in appendix 1. Although the equations for average power output are different for the two tests, the PEP formulas are identical.

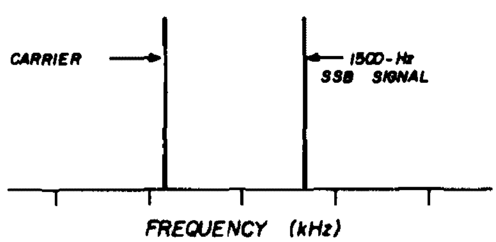
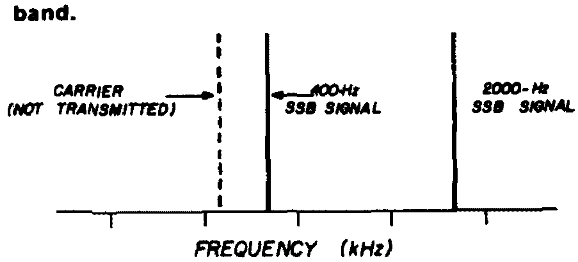


fig. 3. Spectrum of ssb transmitter modulated by 1500-Hz tone and injecting carrier to obtain second rf signal equal in amplitude to the tone.

fig. 2. Spectrum of ssb transmitter modulated by a two-tone test signal containing 400- and 2500-Hz tones and transmitting upper sideband.



multitone relationships

The approximate equations given in appendix 1 are for single- and two-tone conditions, the most common test situations. However, in some multi-channel transmitter applications many more tones are used. The following method can be used to determine the peak-envelope-

power to average-power ratio. (For the purposes of this explanation it is assumed that all the tones are equal.)

The following examples demonstrate two important relationships between single and multitone signals amplified by a linear system.

Assume the amplifier is set up for a single-tone driving signal and a Point "A" on the operating line is established (see fig. 5). A definite PEP output is developed under this condition. To drive this linear amplifier to the same PEP output with a multitone signal, the drive signal voltage for each tone must be 1/nth (n = number of tones) the amplitude of the single-tone signal.

By assuming a perfectly linear amplifier where the input waveshape is exactly reproduced in the output load, these grid waveshapes can be used to demonstrate the relationship of PEP to Average Power.

For the single-tone case, PEP = Average Power; for the two-tone case, PEP = twice Average Power. However, in the two-tone case the average power is one-half that of the single-tone case, so the resultant peak envelope power ratings are identical.*

The two test frequencies (f_1 and f_2) are equal in amplitude but slightly different in frequency. As a result, when they are exactly in phase the two crest voltages add directly to produce the crest of the two-tone envelope. When the two frequencies are exactly out of phase the

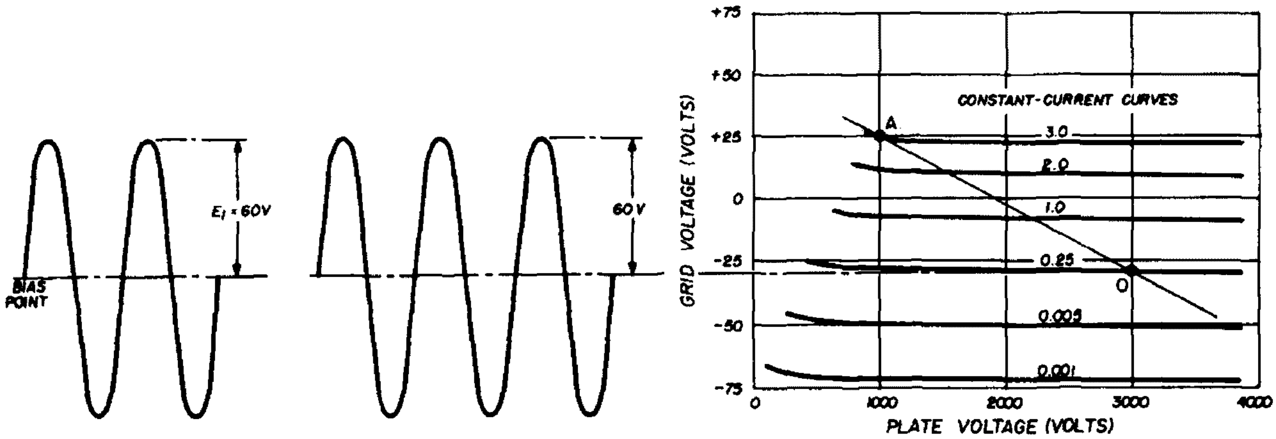


fig. 5. Single-tone condition.

*This is best illustrated with two practical examples.

single-tone

$$\text{Average power} = \frac{E_1(rms)^2}{R_L} = \frac{\left(\frac{60}{\sqrt{2}}\right)^2}{R_L} = \frac{1800}{R_L}\ \text{W}$$

$$\text{PEP} = \frac{E_1(rms)^2}{R_L} = \frac{\left(\frac{60}{\sqrt{2}}\right)^2}{R_L} = \frac{1800}{R_L}\ \text{W}$$

Therefore, PEP = average power

two-tone:

$$\text{Average power} = P_{1avg} + P_{2avg} = \frac{E_1(rms)^2}{R_L} + \frac{E_2(rms)^2}{R_L}$$

$$= \frac{\left(\frac{30}{\sqrt{2}}\right)^2}{R_L} + \frac{\left(\frac{30}{\sqrt{2}}\right)^2}{R_L} = \frac{450}{R_L} + \frac{450}{R_L} = \frac{900}{R_L}\ \text{W}$$

$$\text{PEP} = \frac{(E_{1rms} + E_{2rms})^2}{R_L} = \frac{\left(\frac{30}{\sqrt{2}} + \frac{30}{\sqrt{2}}\right)^2}{R_L}$$

$$= \frac{\left(\frac{60}{\sqrt{2}}\right)^2}{R_L} = \frac{1800}{R_L}\ \text{W}$$

Therefore, PEP = $2 \times P_{avg}$

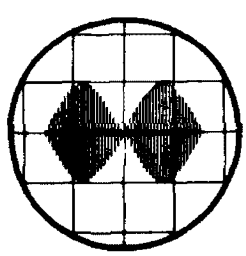


fig. 4. Scope pattern of ssb transmitter modulated by two-tone test signal.

cusp of the two-tone envelope results (see fig. 6).

Note that the voltage amplitude at the crest of the resultant two-tone envelope is equal to that of the single-tone envelope

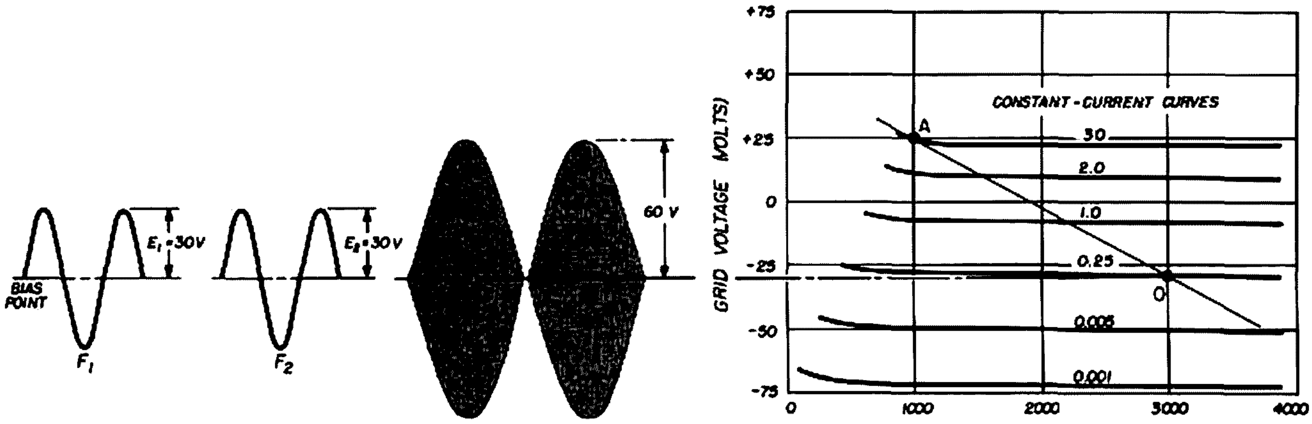


fig. 6. Two-tone condition.

and therefore the tube is driven to the same point on the operating line in each case. If the tube is driven to the same peak plate current and the same peak plate voltage swing by different excitation signals, then the peak envelope power output for both signals is the same.

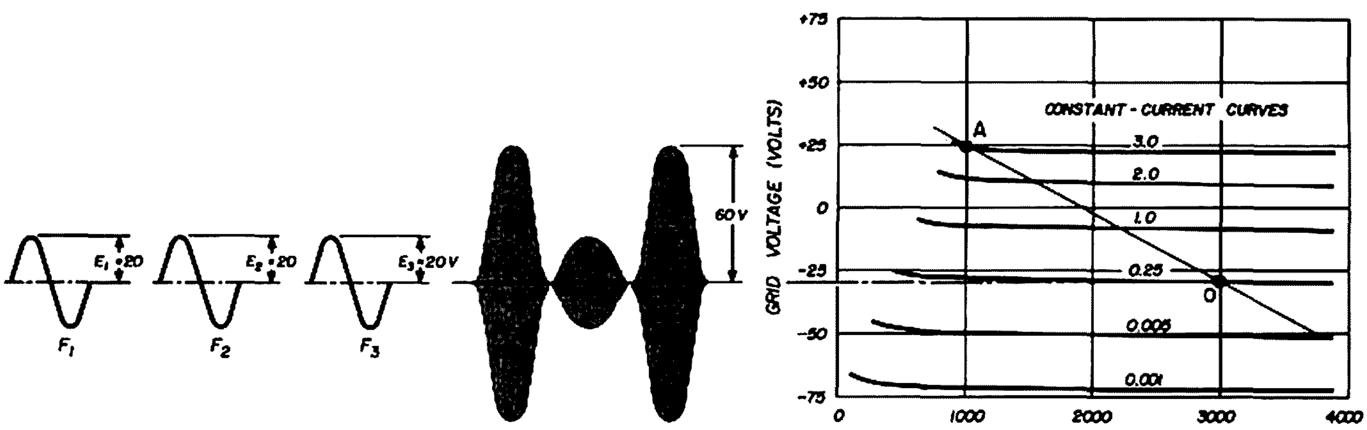


fig. 7. Three-tone condition.

The same holds true for a three-tone test signal. Note that the sum of the three individual tone-crest exciting voltages add in phase to drive the tube to the same peak current and peak plate voltage swing as that of the single-tone case (see fig. 7) so the PEP output is the same as the single- and two-tone examples.

These results (equal amplitude tones) may be summarized by the following expressions:

$$PEP = n\ P_{avg}$$

$$PEP = n^2\ P_t$$

Where P_{avg} is the average power of the composite signal, P_t is the average power in each tone, and n is the number of tones.

example

An fm repeater is to be designed to simultaneously rebroadcast one to eight channels. Each channel must have an average power output of 100 watts. How much peak envelope power must the linear amplifier deliver?

Each channel can be considered to be a single-tone signal. Therefore, the PEP of each channel is equal to the average power of each channel. The *maximum power output* requirement of the amplifier will be under the 8-tone condition. The average power output for the composite 8-tone signal will be 8 times the 100 watts-per channel power. Therefore, the linear amplifier must be capable of 800 watts of average power output.

The peak envelope power will be eight times the average power of the composite signal ($PEP = nP_{avg}$) or 6400 watts. A tube must be selected to deliver this peak-envelope and average power at an intermodulation distortion level compatible with the degree of interchannel crosstalk that can be tolerated.

measurement standards

To describe adequately the performance of a tube in single-sideband linear service, it is necessary to determine many parameters. The normal electrode voltages and currents must be specified as well as the two-tone currents, the operating point, peak envelope power and the magnitude of the intermodulation-distortion products. These parameters are defined as follows:

Peak envelope power is the root-mean-square power at the crest of the envelope. This term is usually shortened to PEP.

Idling plate current determined by the operating point is called the *zero-signal*

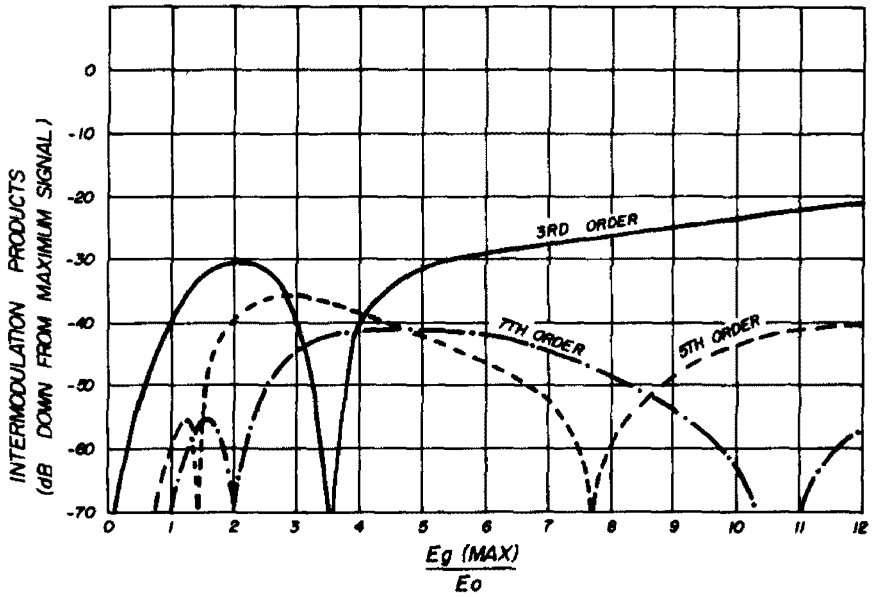


fig. 8. Graph showing intermodulation distortion products. As drive is increased, the various IMD products pass through maxima and minima. Misleading conclusions can be drawn if the equipment is tested near a cusp on the IMD curve where a particular IMD product drops to an extremely low level.

appendix 1

Approximate relationships between meter readings, peak envelope power and average power for class B or AB operation with one- and two-tone tests.

parameter	single-tone	two-tone
dc plate current	$I_b = \frac{I_{pm}}{\pi}$	$I_b = \frac{2I_{pm}}{\pi^2}$
plate input (watts)	$P_{in} = \frac{i_{pm}E_b}{\pi}$	$P_{in} = \frac{2i_{pm}e_p}{\pi^2}$
average output (watts)	$P_o = \frac{i_{pm}e_p}{4}$	$P_o = \frac{i_{pm}e_p}{8}$
PEP (watts)	$P_o = \frac{i_{pm}e_p}{4}$	$P_o = \frac{i_{pm}e_p}{4}$
plate efficiency	$N_p = \frac{\pi e_p}{4E_b}$	$N_p = \left(\frac{\pi}{4}\right)^2 \frac{e_p}{E_b}$

definition of symbols:

i_{pm} = peak of the plate current pulse (plate current pulse is not sinusoidal)

e_p = peak value of plate swing, assumed to be sinusoidal when tank-circuit has sufficiently high Q.

E_b = dc plate supply voltage

plate current and is designed I_{bo}.

The other two plate current values of significance are the *single-tone plate current* and the *two-tone plate current.* The ratio of single- to two-tone current is 1.57:1 in a true class B amplifier (180° plate conduction angle). For other classes of linear operation and for different zero-signal plate currents, this ratio varies from 1.1 to 1.57:1.

The standard method of specifying the magnitude of the distortion products is to specify the reduction in decibels of one product from one tone of a two-equal-tone signal.

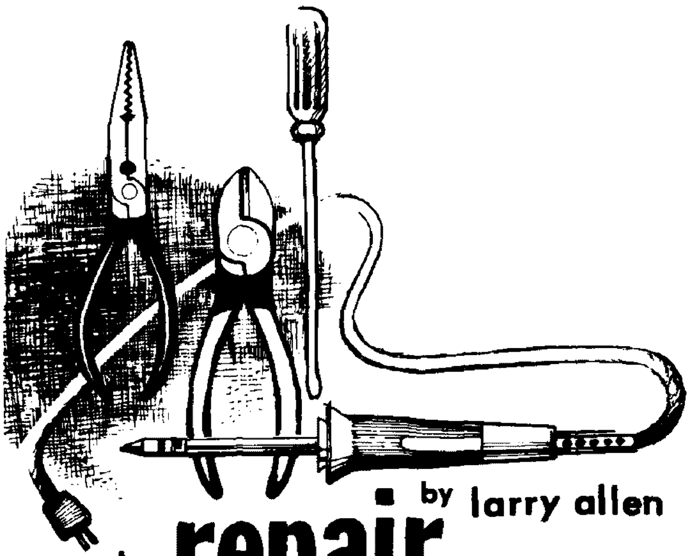
For example, assume that a particular tube under a given set of operating conditions has third-order distortion products of −35 dB and fifth-order distortion products of −50 dB. This means that the third-order product has an amplitude of 35 dB below one of the two test tones and the fifth-order product has an amplitude 50 dB below one of the two test tones. (It is also correct to add the amplitudes of the two third-order products and compare them to the *sum* of the two tones. The decibel ratio is still the same as the example.)

It is *not* correct to compare one distortion product to the sum of the two tones; that is to say, the PEP value of the signal. The resulting distortion figure would be 6 dB better than the correct example (−41 dB rather than −35 dB and −56 dB rather than −50 dB).

Normally the tube under test is adjusted to the full drive condition, and all the pertinent parameters are measured. The drive signal is then reduced. At each test point, all the parameters are measured again. The resulting data can then be plotted as a function of drive voltage.

It should be noted that maximum intermodulation distortion does not necessarily occur at maximum drive level, and it can be shown mathematically that an intermodulation characteristic like fig. 8 can be expected. In practice there is very good correlation between mathematical prediction and actual test results.

ham radio



by larry allen

the repair bench

sweep response curves for low-frequency i-f's

You've been reminding me. Some months ago I promised to show you how to display a sweep response curve for low-frequency i-f stages. This month, I keep that promise.

As you know, test equipment people have quit making sweep instruments that can be used at 60 kHz, 455 kHz, and other low intermediate frequencies. Electronically-swept rf generators of today have two drawbacks at those frequencies. First, their center frequencies seldom go below 4 MHz. Second, their sweep doesn't swing the frequency wide enough at the low end to reach that far below 4 MHz.

What's worse, not all present-day sweep generators even go down that far. Instead, they're designed exclusively for television sweep alignment. A few include a 10.7-MHz center frequency for sweeping i-f strips in commercial fm receivers.

There is a way you can generate a sweep curve centered on ham i-f frequencies. You can do it with a sweep generator that doesn't go near those low frequencies; all you need is an rf generator to use with the sweep instrument.

The secret, as you may have guessed, is heterodyning. Let's review some of the fundamentals of "beat frequencies" and be sure you understand what's happening in the method I'm about to describe.

one and one make four

If you mix two rf signals together in some kind of nonlinear stage, they beat together in a process known as heterodyning. The output of this nonlinear mixer stage is four signals. Two of them are the same as the original signals, one is the sum of their frequencies, and one is the difference.

For example, suppose you *mix* a 10-MHz and a 9-MHz signal in the nonlinear mixer stage. The outputs are: 1 MHz (the difference), 9 MHz, 10 MHz, and 19 MHz (the sum). Which output you use depends on your purpose. In a superhet receiver, you might pick the difference — to form an intermediate frequency. In a sideband transmitter, you might pick the sum — the 19-MHz — to raise the signal to a communications frequency.

For the purpose I'm describing in this *repair bench,* you approach this same principle from another direction. What you need to do is convert a high frequency to whatever is used in the i-f stages. To do that, you mix in a signal whose frequency is displaced from the available sweep signal by exactly the i-f.

Let me cite an example. Suppose the sweep signal is at 10 MHz, and the i-f strip you want to sweep operates at 1 MHz. You can mix a plain rf signal of 9 or 11 MHz with the sweep rf signal. Either frequency produces a difference heterodyne of 1 MHz (1000 kHz). They also produce sum heterodynes, but the i-f tuned circuits reject any but the 1-kHz frequency.

heterodyned sweep signals

If you saw my earlier *repair bench* about sweep alignment in ham gear (April, 1970 issue), you know how a sweep signal is made up. It has a center frequency, which is the frequency at which the sweep generator dial is set.

Here's what this means: A sweep signal that is 1 MHz (1000 kHz) wide swings 500 kHz above and 500 kHz below the center frequency. If the center is 10 MHz, the frequency swings from 9.5 MHz to 10.5 MHz, and does it back and forth 60 times every second.

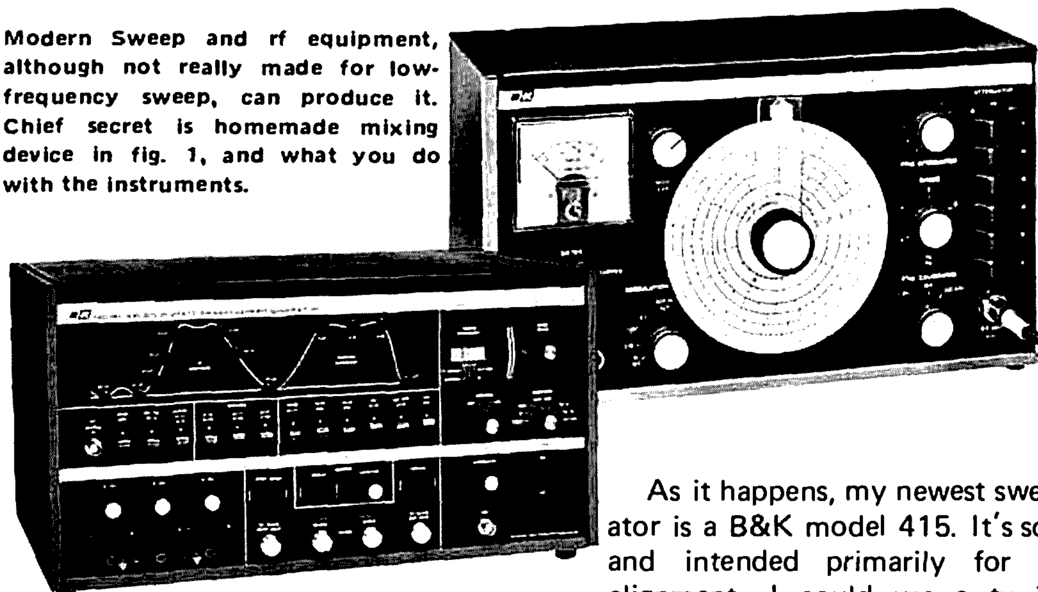
That frequency is swept upward and downward from center. How far up and down depends on the *setting* of the *Sweep Width* control. The sweep *rate* of most generators today is 60 times per second.

The thing about this idea is that you can make the center frequency of the "new" sweep signal whatever you want it to be merely by choosing the unmodulated rf signal you mix with the original sweep signal.

using what you have

You can use almost any sweep generator. Best results come from using the lowest sweep frequency you have available. Just be sure the sweep width is usable.

Modern Sweep and rf equipment, although not really made for low-frequency sweep, can produce it. Chief secret is homemade mixing device in fig. 1, and what you do with the instruments.



Visualize what happens when this same sweep signal is converted to some other frequency. The arithmetic is simple. Imagine it's being heterodyned with an 11-MHz signal. At the instant the sweep signal passes through center frequency at 10 MHz, the difference heterodyne is 1 MHz or 1000 kHz. That's center frequency for the "new" sweep signal. At the instant the sweep signal reaches its bottom frequency, at 9.5 MHz, the difference is 1.5 MHz or 1500 kHz. Then, when the frequency is swept up to its topmost value, at 10.5 MHz, the difference is 0.5 MHz or 500 kHz. So, by mixing a 10-MHz signal, which is swept 500 kHz each way, with a plain unmodulated 11-MHz rf signal, you generate a 1000-kHz signal swept 500 kHz up and 500 kHz down.

As it happens, my newest sweep generator is a B&K model 415. It's solid state, and intended primarily for television alignment. I could use a tv i-f sweep frequency, in the 44-MHz range. However, this generator also produces a sweep at 10.7 MHz, for aligning the i-f strips of fm receivers. And there's a nice marker right in the middle – precisely at 10.7 MHz. (More about markers later.) That's lower than the tv sweep frequencies, so I use the 10.7 MHz for heterodyning down to low sweep frequencies.

The source of unmodulated rf at my bench is a new B&K model E-200D. It's a modern version of the Old Precision Apparatus model E-200. Old-timers will remember that once-popular unit. The dial settings are reasonably accurate. But more important, the E-200D is stable once it's warmed up. That's a requirement of whatever generator you use.

The calculations for using this pair of instruments are easy. They apply to any

sweep and rf generators you use. To create the low-frequency sweep I want from the fixed 10.7-MHz sweep signal, I either subtract or add the frequency of the i-f strip I'm checking.

A common i-f is 455 kHz. That's what's in the receiver I'm using for this demonstration. With a 10.7-MHz sweep signal, a 455-kHz beat can be obtained by adding 11.155 MHz or 10.245 MHz. Either one is okay, but I choose 10.245.

If the i-f were 60 kHz, another common one in ham receivers, an rf signal at 10.64 MHz would do the job.

You have to build a little mixer stage yourself. The simple one in fig. 1 is adequate. The diode is the nonlinear device that produces the heterodyning between the two signals.

Of course, you need a scope. Any ordinary servicing scope will do. Mine is a modern wideband scope, but for this purpose yours needn't be.

The model 415 sweep/marker generator uses post-injection for the marker, as most recent instruments do. The block diagram in fig. 2 shows the equipment hookup. Here are the steps:

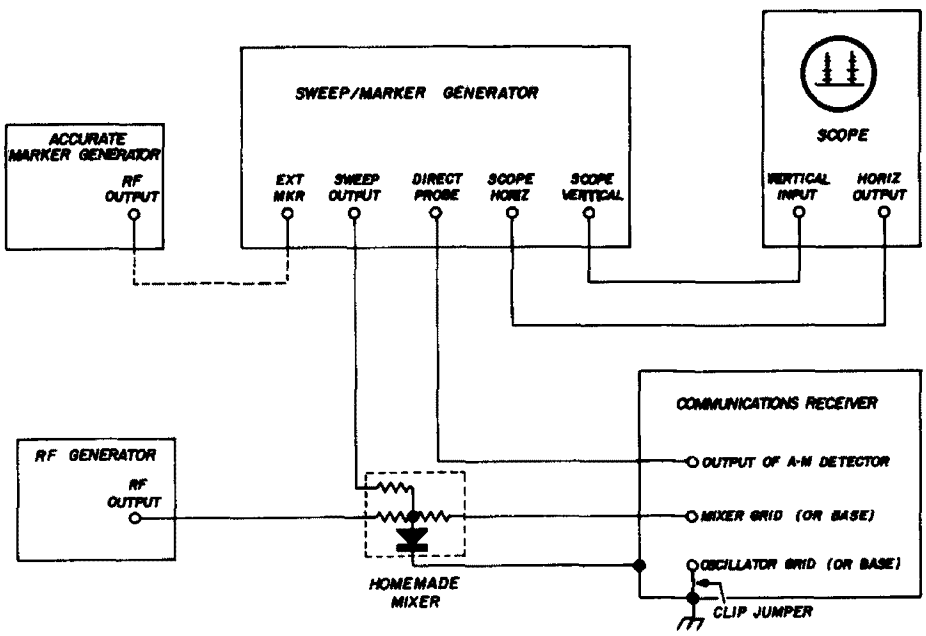


fig. 2. Equipment hookup. It is slightly different if sweep generator doesn't have the post-injection marker arrangement.

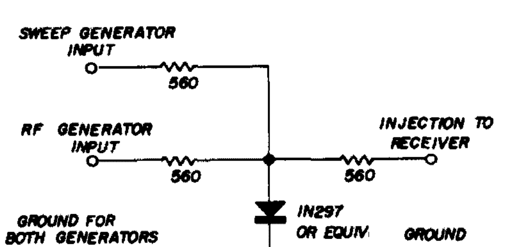


fig. 1. Four parts make up mixing device used for creating low-frequency i-f sweep signal.

1. Let everything warm up at least 15 minutes.
2. Set scope sweep to External Input.
3. Set scope vertical-input attenuator to whatever setting gives a 1-inch deflection for about 5 volts of signal input.
4. Connect scope vertical input to Scope Vertical jack of sweep/marker generator.
5. Connect scope horizontal input to Scope Horizontal jack of sweep/marker generator.
6. Connect rf outputs of the two generators to the inputs of your homemade mixer device. Connect ground cables to receiver chassis.

7. In the receiver, clip the mixer-device diode lead to ground and the output lead to the grid (or base) of the receiver's mixer stage.
8. Clip a shorting jumper from the receiver oscillator grid (or base) to ground (or to emitter), thus disabling the receiver's oscillator.
9. Connect the direct-probe lead from the sweep/marker generator to the output of the receiver's a-m detector. Note: *If the receiver is multiple-conversion*, having more than one i-f section, make a diode detector to use at the output of each i-f strip and display the curve for only one i-f section at a time.

That's how the equipment hookup goes. Be sure all grounds are in place. Likewise doublecheck the connections. False curves can be the result of careless hooking-up.

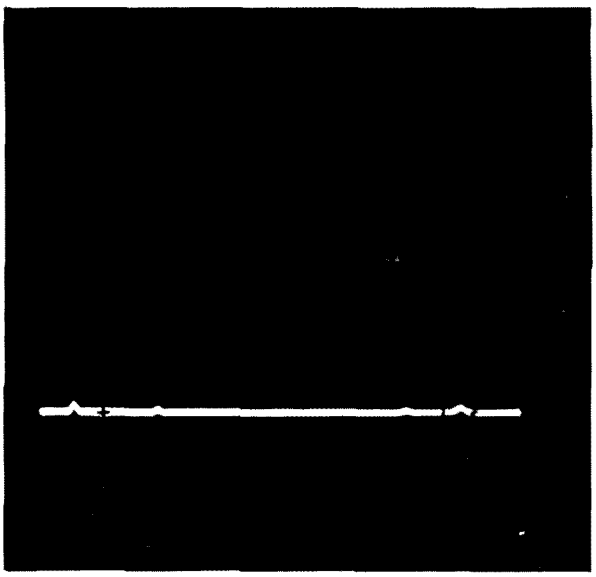
getting the curve

The 10.7-MHz sweep in the model 415 has a center-frequency marker. That offers a convenient way of situating the sweep just right on the scope trace.

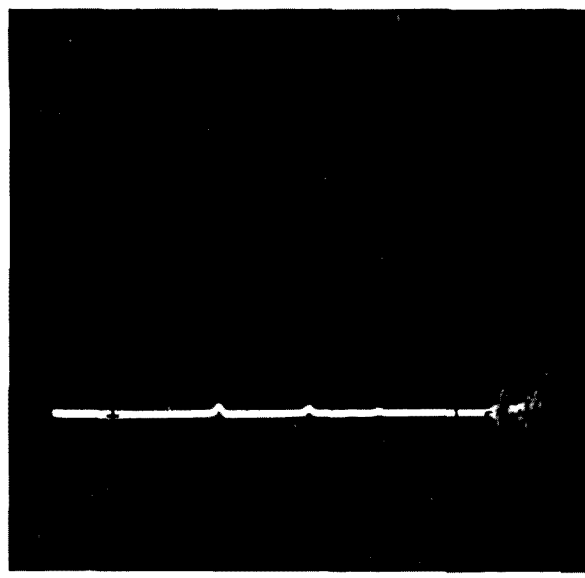
Here's how. Leave the rf generator output control at zero for now. Turn the sweep rf output wide open. Set the *Sweep Width* control as low as it goes; it'll still be wide enough for communications receiver i-f coils. Turn up the *Marker Amplitude* control on the generator just slightly.

Now roll the *Center Frequency* dial back and forth until the marker comes into view along the scope trace. The photo in fig. **3A** shows the trace before you get the center frequency just right. This is sort of a fine-tuning control. **Fig. 3B** shows the marker just coming into

A



B



C

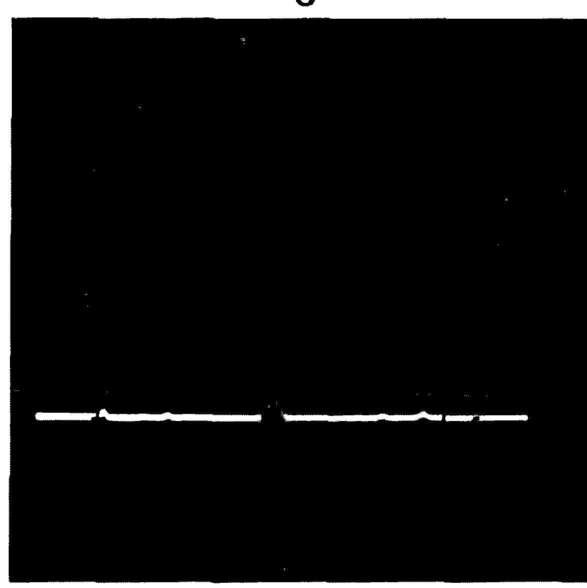


fig. 3. Traces you see when first setting up for low-frequency sweep of receiver i-f section. A shows trace without curve or marker in view. B is marker with frequency slightly high. C shows marker (and frequency) centered.

view. Turn the *Center Frequency* control to put the marker exactly in the center of the scope trace, as in **fig. 3C**.

The object of this step is to put the marker in a position that enables you to use it as the center of the curve you will presently display. The marker is at 10.7 MHz, but it is added to the scope trace from within the sweep generator. It therefore remains as a center reference for the 455-kHz sweep signal developed when the rf signal is added.

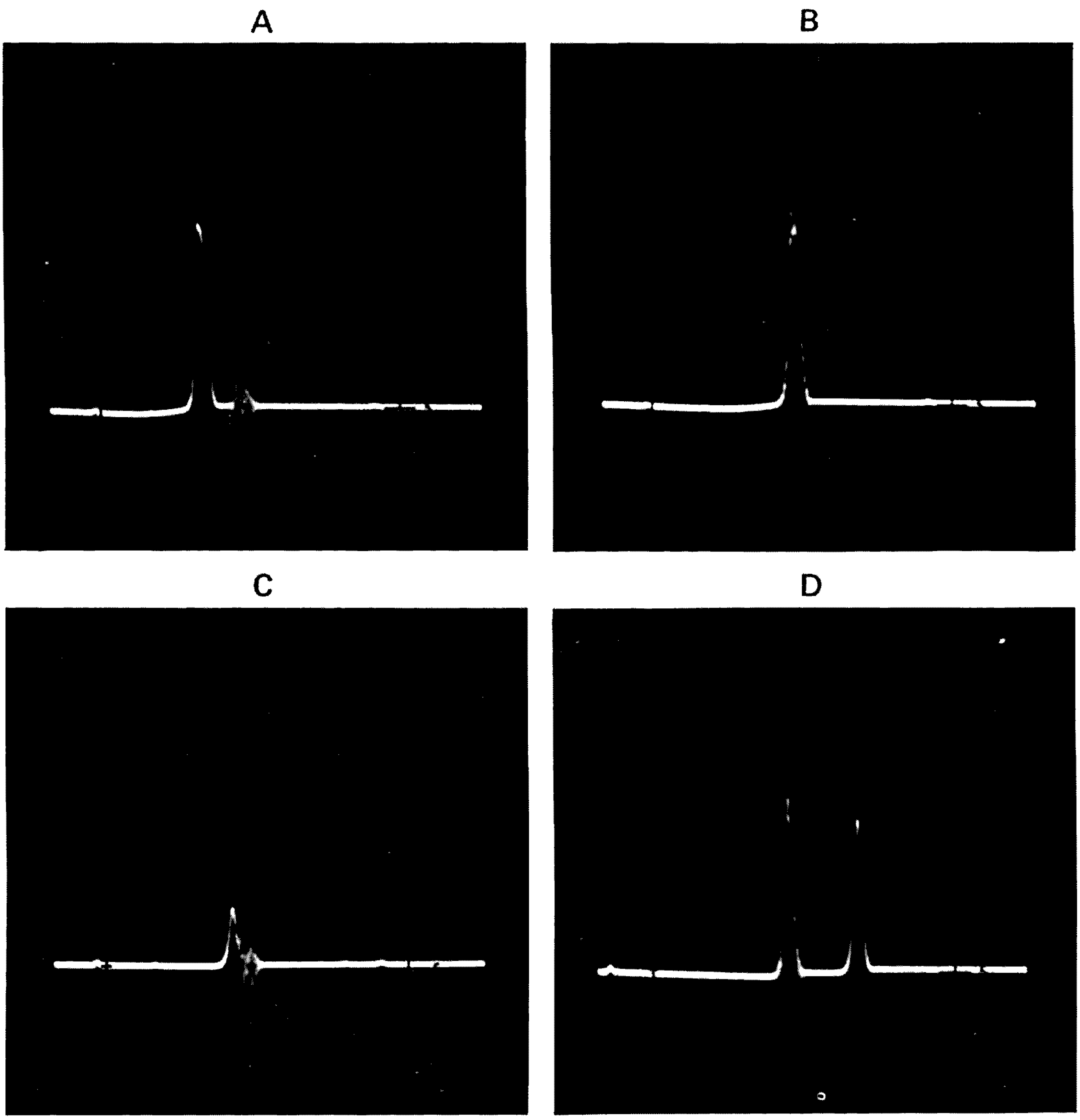


fig. 4. Curve appears when rf generator signal is added. Curve in A not exactly at i-f center marker. Curve in B moved to coincide with marker by returning i-f coils or rf generator. In C some of i-f coils detuned. Two curves in D mean the receiver oscillator hasn't been disabled.

Now set the frequency dial of the rf generator as near 10.245 MHz as you can. Turn up the output controls. If your hookup is correct and the i-f strip is reasonably normal, you'll see the curve rise somewhere along the scope trace, as in **fig. 4A**. Just how far it is from the center marker depends on how near the rf generator is to 10.245 MHz and how accurately tuned the i-f coils are.

(The ideal is a *calibrated* 10.245-MHz signal, such as from a heterodyne frequency meter that has been checked against WWV. I've used a BC-221 for this, as well as a Lampkin model MFM frequency meter. For this demonstration, I calibrated the vernier of my rf generator against a frequency meter. That way, I know the i-f stages are slightly off-frequency when I see the curve displaced from the marker as it is in fig 4A.)

The photo in fig. 4B shows the curve right on the marker. You can put it there by adjusting the dial of the rf generator. Since I knew my rf generator was precisely accurate, I readjusted the six i-f coil cores (two per i-f can) to move the curve over to the marker.

The technique is not difficult. I just turn each core a small amount in the same direction, then peak all six at the new position of the curve. The curve in fig. 4B results. The curve in fig. 4A was down-frequency from the marker. I therefore backed all six cores out about one-half turn. At first, it looked as though I was destroying the curve. (Fig. 4C shows how it looked when only three slugs were turned to the new setting.) But as all six were turned near the new frequency, the curve got tall again.

Be sure the receiver oscillator is cut off. I told you how, in step 8 of the equipment hookup. Otherwise, you may get the double curve shown in fig. 4D. The correct curve is identified by the marker, but it's best not to have the extra curve.

You may wonder how to determine bandwidth. With another generator of high accuracy, you can put a movable marker into the sweep generator through the *External Marker* jack. Record the frequencies when the marker is at the upper skirt (fig. 5A) and at the lower skirt (fig. 5B). Bandwidth is roughly the difference between them. The two markers in fig. 5 are at 10.235 and 10.255 MHz. The difference is 20 kHz, the bandwidth at the bottom of the skirts. That means bandwidth at the half-points on the curve is 10-12 kHz. That's not bad for a general-purpose receiver, but a little wide for a selective communications receiver.

Another way I estimate bandwidth is with the built-in 100-kHz markers that are part of my model 415 sweep generator. Fig. 6 shows the trace with the 100-kHz markers activated. You may have to increase *Sweep Width* slightly to get both markers on the scope trace.

A B

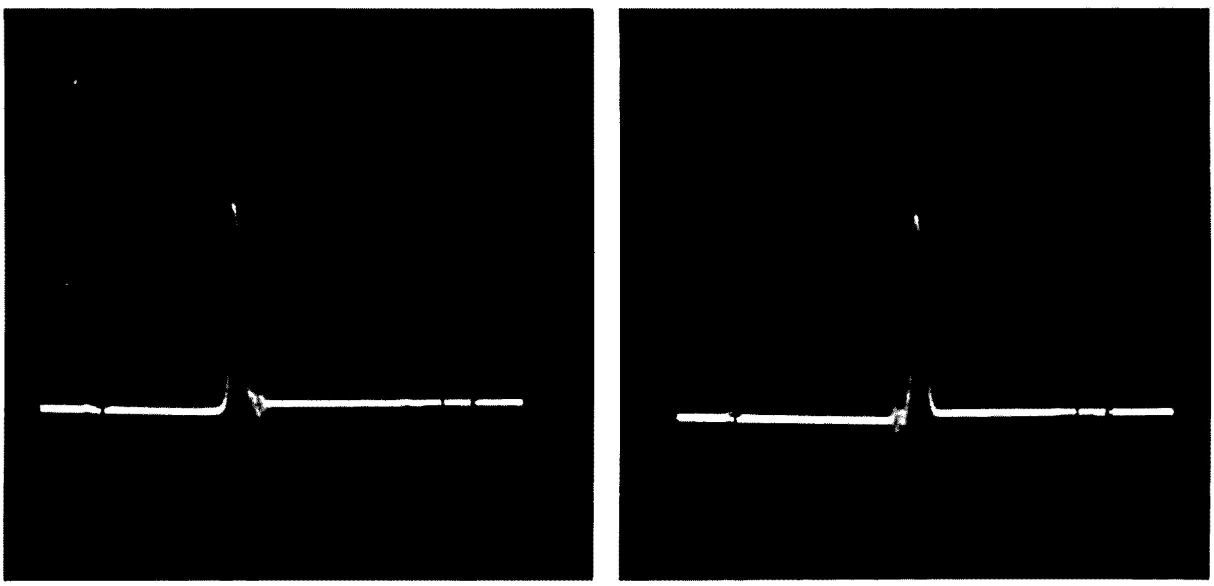
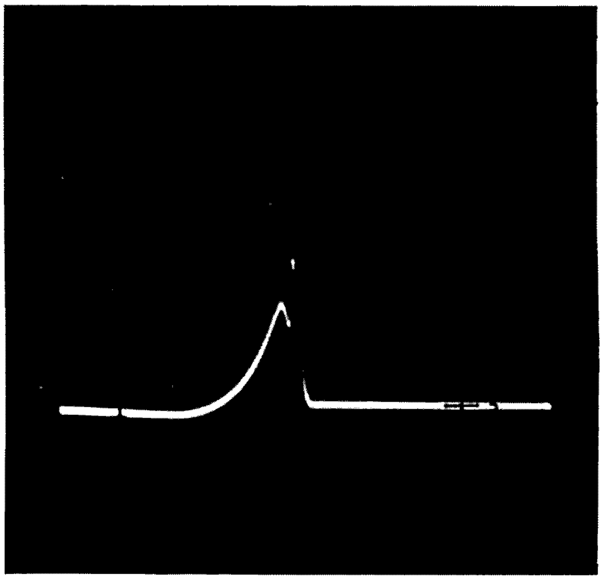


fig. 5. Movable markers show bandwidth in terms of kHz between upper skirt A, and lower skirt, B.

The 100 kHz beats with the 10.7-MHz marker and puts one at 10.6 and another 10.8 MHz. Converted in the homemade mixer, this means markers at 355 kHz and 555 kHz. Of course the main 10.7-MHz marker still makes the 455-kHz center marker. The distance between the two outside markers is 200 kHz. From that you can estimate the bandwidth of the curve being displayed.

A



B

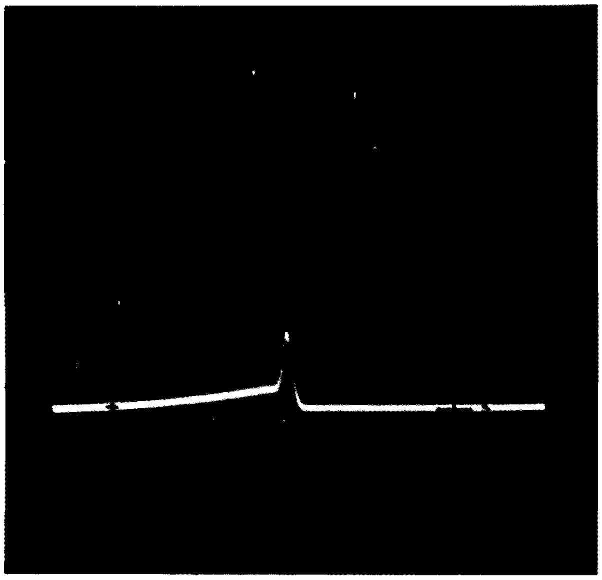


fig. 7. Selectivity switch of receiver changes response of i-f strip, as shown by the curve. Broad response, A, with crystal phasing set at one spot. Sharp crystal position, B, reduces gain but sharpens curve peaking.

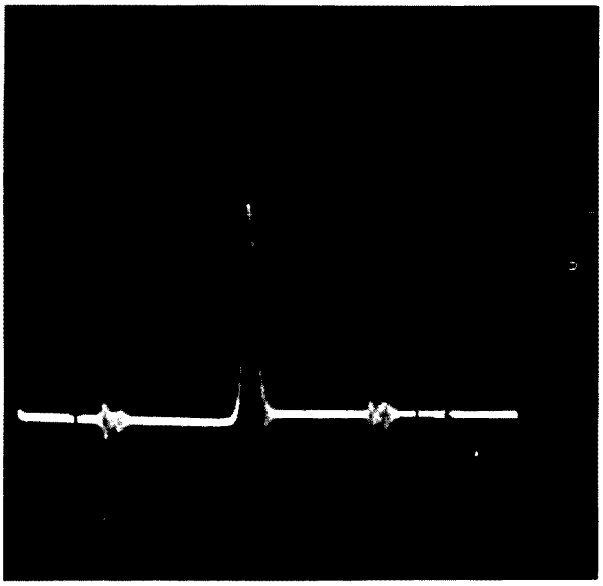


fig. 6. 100-kHz markers on each side of main center marker also can help estimate bandwidth.

other uses of sweep curve

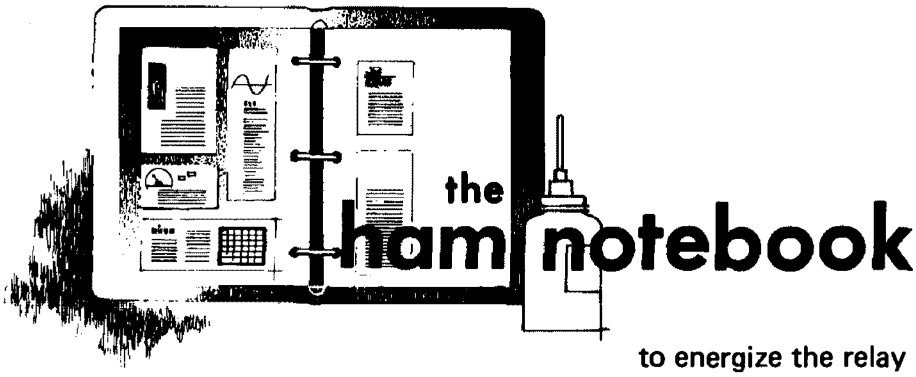
You can see the effect of a crystal selectivity filter on the i-f response. **Fig. 7A** shows the curve with the *Selectivity* switch of the receiver set for *Crystal – Broad*. Turning the *Crystal Phasing* capacitor knob changes the curve shape somewhat.

Switching to *Crystal – Sharp* reduces the gain of the i-f strip, but sharpens the curve. You can see the result in **fig. 7B.** You can alter the *Crystal Phasing* to virtually eliminate the i-f curve.

A sweep curve generated as described here can be used for testing any rf or i-f stages. Without the gain provided by amps in the receiver, you'll need a more sensitive setting of the scope's vertical input. Or, you can build an untuned transistor amp to follow the homemade mixer. That'll give you enough signal to be useful with unamplified tuned circuits. You'll have to add a diode detector for stages that don't end with an a-m detector.

With a little practice and imagination, you can find a lot of other uses for this method of generating a low-frequency sweep signal.

ham radio



safer suicide cord

Most electronics experimenters have an ac-plug-to-clip-lead adapter (better known as a "suicide cord") on their workbench. Unfortunately, too many experimenters fail to take a few simple safety precautions – precautions that can save much embarrassment or worse!

First of all, always use a fused ac plug such as the El Menco along with 5- to 10-amp fast-blow fuses. For the cord itself use about 6 feet of no. 16 heavy-duty lamp cord. For the clip-lead end, Mueller no. 60 alligator clips are a good bet, but be sure to insulate each alligator clip with a Mueller no. 62 rubber cover.

Bruce Clark, K6JYO

using undervoltage relays

Although many forms of undervoltage relays are on the market, it sometimes happens (as in my case) that a quick and dirty substitute can be found. Most standard relays exhibit a very wide difference in pickup and dropout current, typically three to one. Therefore, it's often impossible to use a single relay for a specific application.

I needed a relay that had to drop out within approximately 10% of the normal running voltage. Available relays were Allied Control type BO6-D 184VDC. These units have a 13500-ohm coil. Pickup is at approximately 10 mA, but dropout occurs at about 3 mA.

As shown in **fig. 1**, R1 is proportioned to energize the relay (RY 1) at about 20% below the operating voltage. R4 is then set to pull in RY 2 at a point a little higher – say 10% below the operating voltage. The 100k composition pot is then set to drop out RY 1 at a point 10% below the operating voltage.

This arrangement was run in a bench setup for eight hours to determine if relay heating would result in any variation of the dropout point. The maximum variation in dropout point was approximately 2% of the operating voltage.

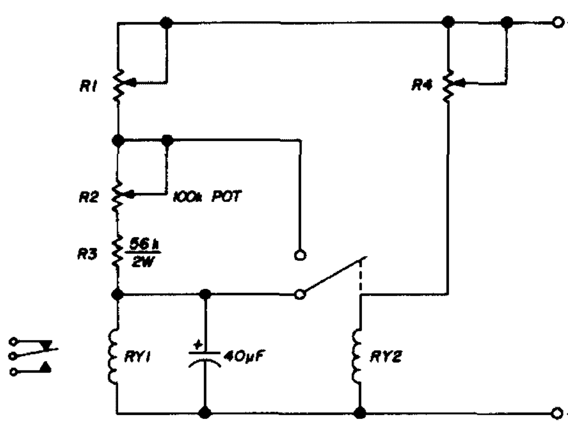


fig. 1. Method for controlling relay dropout voltage. Resistors can be proportioned to cause pull-in and dropout over fairly wide current range.

caution

Don't try to use a variation in dropping resistance as shown in **fig. 2**, or instability will result. Also, don't try to use an extra set of contacts on RY 1 to perform the function of contacts on RY 2; i. e., to switch in R2 and R3. Oscillation or instability will result.

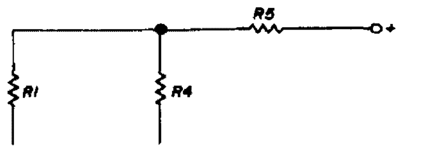


fig. 2. An example of dropping-resistor arrangement that will cause instability.

A solid-state circuit would have been preferable, except that this particular equipment was to be subjected to a radioactive environment, which precluded the use of solid-state devices.

Bill Wildenhein, W8YFB

simple resistance standard

As many amateurs know, ohmmeter readings taken with a vom can be notoriously inaccurate. However, there are times when you need an accurate resistance measurement of a higher order than that obtainable with an ordinary multimeter. Higher accuracy generally requires a Wheatstone bridge or similar laboratory-type equipment. For occasional use such expense cannot be justified.

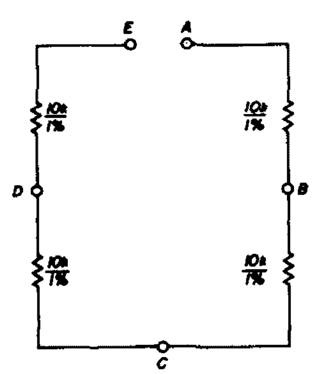


fig. 3. Simple resistance standard provides accurate checkpoints from 2500 to 40,000 ohms. Connections are shown in table 1.

The inexpensive "resistance checker" shown in the photograph and in fig. 3 is a simple and direct solution to this problem. Four 1% resistors are wired as shown; it is apparent that there will be resistance check points at 10k, 20k, 30k and 40k ohms. If your volt-ohmmeter has not been grossly abused, you can set the

table 1. Parallel connections required for resistanca values from 2500 to 40 kilohms.

desired resistance value	connect together	read ohms between
2500	A-C B-D C-E	A-B
3333	A-C B-D	A-D
5000	A-C	A-B
6667	A-D	A-C
7500	A-E	A-D
10,000	—	A-B
13,333	A-C B-D	A-E
15,000	A-C	B-D
16,667	A-D	C-E
20,000	—	A-C
25,000	A-C	B-E
30,000	—	A-D
40,000	—	A-E

meter to 10k ohms when connected across one leg of the standard — merely adjust the "zero ohms" knob so the meter coincides with the selected resistance value. Thereafter you may read resistance values with good accuracy up to 10 or 20 percent removed from the check point.

In addition to the multiple 10k check points, parallel connections will result in resistor standards at 2,500, 3,333, and 5,000 ohms (see table 1). Other standard values may be obtained through suitable interconnection of the various resistors. The chart is provided since most of us don't want to get out the slide rule when a quick resistance check is needed. Obviously other decade values may be selected but the above setup provides resistance checkpoints at more than one dozen intervals throughout the range from 2.5k to 40k ohms.

Neil Johnson, W2OLU

cleaning printed-circuit boards

If you work with printed-circuit boards you probably have had difficulty removing solder from the holes when removing resistors, capacitors or other components. A cutting-torch cleaner — available from a welder's supply house — is a very useful aid. The cutting-torch cleaner comes with 12 different-sized steel rods which can be pushed through the solder holes. About one-quarter inch back from the smooth tip of each rod there's a machine-cut rasp surface which may be used to enlarge the holes, if necessary. The rods are usually furnished with a small metal case for easy storage.

Felix W. Mullings, W5BVF

Heath HW-17 modifications

After using the Heath HW-17 for some time, I decided that the receiver in the little rig lacked sufficient gain. The frontend seemed quiet and relatively sensitive, but overall receiver response was low. When WB2EGZ's article appeared in the August issue I built the mosfet preamplifier. It did give the receiver more gain, but didn't help much with weak-signal reception. In fact, the preamp seemed to degrade the noise figure of the HW-17's own frontend.

Since I already had the 40673 transistor and its bias and load resistors, it was decided to try to use those parts if possible. Since the frontend seemed to be quiet and sensitive I thought that the receiver needed a little more i-f gain. From a quick check of the diagram it appeared that the 40673 mosfet and associated components would fit the HW-17 like a glove. It was wired into the set between the i-f output and the second mixer. This can be done very easily without drilling any holes.

Solder is removed from the i-f output pin of the tuner where it connects to the board; a vacuum de-soldering tool is very handy here. A small piece of spaghetti is slipped over this pin to keep it from touching the foil. Gate 1 of the transistor is soldered directly to the pin, and the drain lead is soldered to the foil nearby. The transistor is mounted perpendicular to the board and the rest of the components are soldered to the transistor leads. Although the additional i-f stage is very stable, and no combination of lead placement will cause it to oscillate, use short leads.

The modification requires re-alignment of T1 (in the tuner) and L6. Adjust them for maximum signal strength as indicated on the S-meter while rocking the main tuning dial across a weak signal.

Since adding this modification, the familiar background hiss of most vhf receivers is present, and ignition and atmospheric noise can be easily heard. With the antenna disconnected, the noise is very low, indicating the frontend is quite quiet. The S-meter is much livelier because of improved agc action. Also the squelch will open with weaker signals than before.

Ed Ranson, WA5PWX

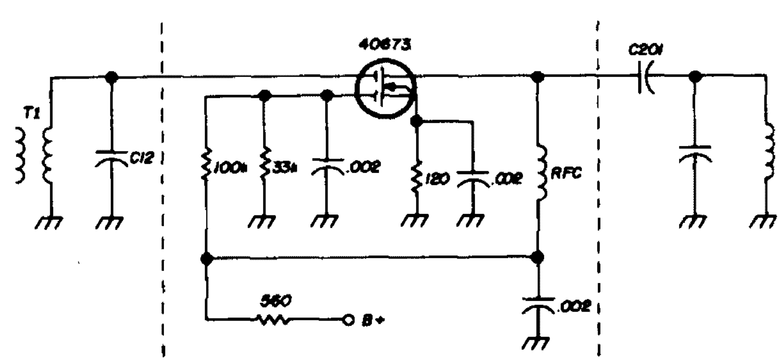


fig. 4. Added stage of i-f amplification for the HW-17. RFC is not critical.

noise limiter for heathkits

Since I live in a noisy area and own one transceiver with a noise limiter and one without, the advantages of a noise limiter are very apparent to me. With some component value changes to match the resistance of the original audio-gain control in my transceiver, a very satisfactory noise-limiter circuit evolved for the Heathkit SSB line. The components for this circuit cost less than $3.00, and the time involved is approximately one hour. The circuit is shown in fig. 5.

The switch on my HW-12 is mounted as low as possible and centered on the front panel. I used a miniature switch that mounts with two screws. These screws are used to support two 4-connector tie points.

If the af gain-control in your transceiver is a 500k pot, R1 and R2 will be 1 megohm and R3, 150k. If the af gain-control is a 1-megohm pot, then R1 and R2 will be a 2.2 megohm, and R3, 270k. All resistors are 1/2 watt.

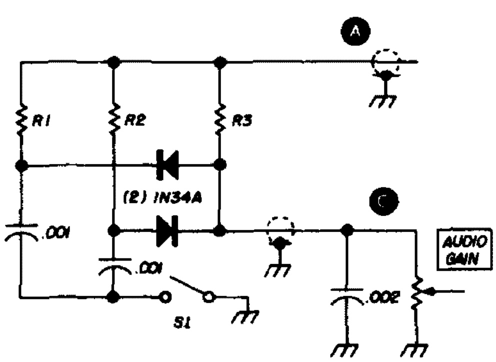


fig. 5. Simple noise limiter for Heathkit ssb transceivers. Values for R1, R2 and R3 are given in text.

Remove the shielded audio-input lead from point C of the af gain-control and connect it to the input of noise limiter at point A. Connect the output of the limiter to point C on the af gain control. You will find there is very little loss with this limiter and no noticeable distortion until the audio is too loud for comfortable listening.

Jim Welborn, W7CKH

increasing the versatility of the comdel speech processor

The Comdel speech processor can be made more versatile by the simple modification shown in the diagram. A miniature potentiometer is mounted on the rear plate of the unit and connected into the output circuit so the output level may be

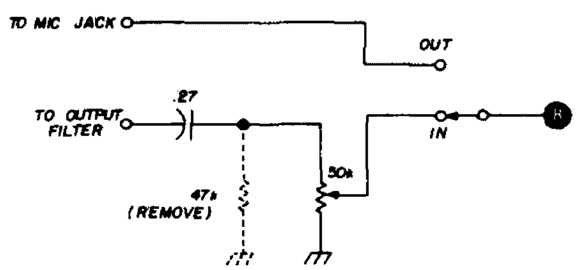


fig. 6. Gain control for Comdel CSP-11 adds versatility to the unit.

adjusted to be equal in peak amplitude to the level encountered when the unit is switched out of the circuit.*

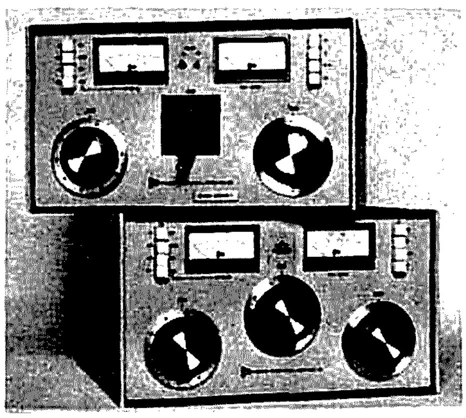
Gain through the speech processor, without the potentiometer, is greater than unity and "in-out" checks can only be made with difficulty because of change in gain level. With the addition of the potentiometer, processor output level may quickly and easily be set to the proper amount. The output circuit potentiometer is substituted for a 47k resistor (shown in dotted lines) which is removed from the circuit board.

William I. Orr, W6SAI

*With high-level microphones, such as the D104, the peak microphone output is considerably higher than the limited 70-to 100-mV peak level available from the processor; in this case the output potentiometer should be mounted in the microphone line.

new products

eto linear amplifiers



The new Delta 70 linear amplifier is externally identical to the Alpha 70 except for the plate-tuning control; it uses a vernier drive to an air-variable capacitor. Internally, the only difference between the Delta and Alpha models is the use of an air-cooled 3-1000Z in place of the Alpha 70's vapor-cooled system. The two amplifiers are almost identical electrically and handle much the same. Power ratings are identical.

While the Alpha 70 features continuous coverage from 3 to 30 MHz, the Delta is specified for the amateur bands, 10 through 80, although it too is almost continuous tuning. The Delta 70 linear amplifier is initially priced at $1195.00.

For more information on the Alpha 70 or Delta 70 linear amplifiers, write to Ehrhorn Technological Operations, Inc., Post Office Box 1297, Highway 50 East, Brooksville, Florida 33512, or use *check-off* on page 94.

mosley 75 and 80-meter conversion kit

The new Mosley RV-8C conversion kit is designed for amateurs who already own the Mosley 4-band vertical, the RV-4C, or for those who are considering the purchase of a 5-band vertical antenna system for operation on 10 through 80 meters. The 5-band *antenna stands* only 22-feet high and requires no guying or concrete footing.

The conversion kit includes a loading coil rated at 750 watts a-m or CW, and 2000 watts PEP ssb. Tuning is accomplished by sliding the U-shaped matching-section along the vertical element until a match is obtained at the desired operating frequency.

A plexiglass coil housing, closed at one end with a polyethylene cap, assures protection against the most severe weather conditions. Easy-to-follow instructions are included with the kit. For more information, write to Mosley Electronics, Inc., 4610 N. Lindbergh Boulevard, Bridgeton, Missouri 63044, or use *check-off* on page 94.

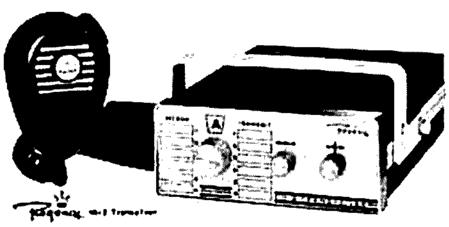
50-watt mobile amplifier

Varitronics has just announced a new power amplifier that should appeal to all mobile fm operators.

The Varitronics PA-50A is a completely solid-state class-C rf amplifier designed for use in mobile amateur fm applications. Because of internal rf

switching, the PA-50A can be used with any two-meter transmitter with 5 to 15 watts output when used with the IC-2F transceiver (nominal 12-watts drive). Balanced-emitter semi-conductors insure complete insensitivity to high vswr – even no load conditions. This handsome and ruggedly built amplifier is styled like the Varitronics IC-2F, features a calibrated output meter and comes complete with a mobile-mounting bracket and dc power cord. $129.95 amateur net. For more information, use *check-off* on page 94, or write to Varitronics Incorporated, 2321 East University Drive, Phoenix, Arizona 20665.

two-meter fm transceiver



Regency Electronics has announced a compact mobile fm transceiver for operation in the 144-MHz band. The solid-state model HR-2 features a 10-watt power output with operation on any of six transmit and receive channels in the band. Simple operator modification, however, can enable the radio to transmit and receive on any of 12 different duplex combinations.

The receiver is a double-conversion superhetrodyne with a highly selective ceramic filter for operation on both wide- and narrow-band signals. Sensitivity is rated at 0.35 μV, 20 dB quieting and audio output at 5 watts.

The transmitter features phase modulation for exacting carrier stability. Individual trimmer capacitors enable frequency setting optimum performance in point-to-point or repeater applications. Built-in swr load mismatch circuitry protects against open or shorted antenna conditions.

The modern integrated circuitry operates on 13 Vdc with low current drain. The package comes complete with plug-in ceramic microphone, built-in speaker and mobile-mounting bracket. One pair of factory installed transmit and receive crystals on 146.94 MHz are included in the $229.00 amateur net price. An ac power supply and linear amplifier will soon be available as optional equipment.

For more information, write to Regency Electronics, Inc., 7900 Pendleton Pike, Indianapolis, Indiana 46226, or use *check-off* on page 94.

short-wave listening

A new, well-illustrated, multi-page primer on short-wave listening has been written by The Hallicrafters Company, and is being offered free to the interested reader, DXer, short-wave listener, hi-fi buff, hobbyist and amateur radio enthusiast.

Aptly titled "short-wave puts you where it's at," the multi-colored brochure takes the interested reader through the many adventures he can experience with short-wave; explains graphically and in simple terms "what is short-wave;" tells the reader what can be heard on short-wave (i.e., amateur bands, overseas marine radio, aeronautical bands, international broadcasting, military and clandestine stations, standard time signals, public safety and many others); provides a complete radio frequency spectrum in a unique and copyrighted graphic illustration; and points out what to look for when choosing a quality short-wave receiver.

Also included in the center fold of the primer is Hallicrafters' complete line of a-m and fm general-coverage short-wave receivers, special frequency monitor radios and amateur transceivers and accessories. For a free copy of the multi-page, short-wave primer, "short-wave puts you where it's at," write to The Hallicrafters Company, Dept. PR, 600 Hicks Road, Rolling Meadows, Illinois 60008, or use *check off* on page 94.

new allied catalog

Allied Electronics' new 1971 catalog of electronics parts and supplies, considered by many as the bible of the industry, is now available. The catalog lists over 70,000 separate stock items from more than 700 manufacturers. Detailed specifications, descriptions and illustrations cover a vast array of parts and components including: semiconductors, integrated circuits, tubes, relays, timers, transformers, resistors, capacitors, connectors, coils, chokes, sockets, plugs, jacks, switches, fuses, batteries, clips, lamps, wire and cable, and much more.

Other major sections of the catalog feature test instruments, recording equipment, sound equipment, intercoms and business communications gear, power supplies, electronic counters, chemicals, hardware, technical books, tools and solder equipment. Allied Electronics Catalog No. 710 sells for $2.00 (or no charge with order) and is available from Allied Electronics, 2400 W. Washington Boulevard, Chicago, Illinois 60612.

solid-state rtty demodulator

The new Space One RTTY demodulator from J&J Electronics has a number of interesting features, including heavy-duty loop supply, regulated power supply, electronic keyer stage supplying plus and minus voltages with provisions for adjusting to 1 mA when driving diodes, as well as both 850 and 170 shifts, selectable from the front panel. The solid-stage circuitry uses integrated-circuit amplifiers

and high-voltage transistors for keying the loop supply.

The Space One uses standard tones of 2125 and 2975 Hz, has metered tuning (indicator lamps available as option), and provides a spare socket for the addition of a three-pole Butterworth bandpass input filter if local conditions warrant it. The basic unit is completely self contained, including loop supply and fsk driver, and is ready to connect between your receiver and printer.

Several extra features are available at optional extra cost, including receive and standby indicator lamps, auto receive without motor-control relay, and auto-start plus motor control relay. Basic unit, less options, is $124.95. For more information, write to J&J Electronics, Windham Road, Canterbury, Connecticut 06331, or use *check-off* on page 94.

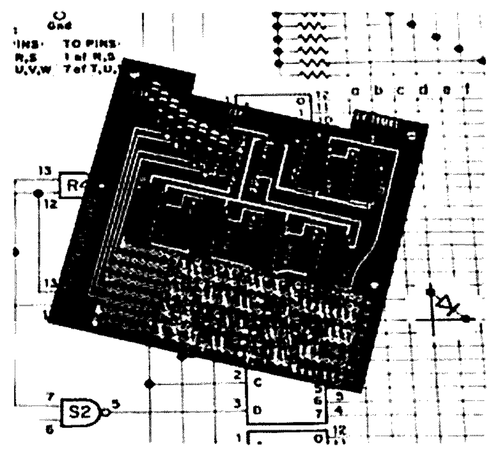
video/rf selector-control center

The new Alco CCV3 is a three-position control center that has many uses in video or rf layouts. Although it is suitable for switching *low-power* antenna systems it is designed primarily for controlling tv cameras and monitors. The unique push ON and push to RELEASE switches allow the operator to select equipment optionally at will. The unit uses standard type-UHF coaxial connectors, and no external power is required for operation. Crosstalk between adjacent channels is negligible. Price is $19.95. For more information use *checkoff* on page 94, or write to Alcoswitch Division of Alco Electronic Products, Inc., Post Office Box 1348, Lawrence, Massachusetts 01842.

two-meter fm antenna

Varitronics has recently introduced the *Redhead,* a two-meter groundplane antenna with 3.4-dB gain. This sturdily built, commercial quality antenna is adjustable for low swr over the entire two-meter band. Radials and radiator are made from heavy-guage aluminum tubing, and the loading coil is well protected by a metal shield which is painted red (hence the nickname). The model AS-2HG *Redhead* antenna is rated at 250 watts cw, a-m or fm, and 500 watts PEP ssb. $18.95 from your local dealer. For more information, write to Varitronics Incorporated, Post Office Box 20665, Phoenix, Arizona 85036, or use *check-off* on page 94.

keyer memory kit



For hams who like to roll their own, Curtis Electro Devices has introduced a do-it-yourself diode-matrix message memory for their EK-39M, a combined keyer and message generator. The kit is an exact plug-in equivalent to the custom integrated-circuit memory which Curtis supplies individually pre-programmed per user instructions. It consists of six integrated circuits, a pre-drilled circuit board and 150 diodes plus other components and hardware. Complete instructions are provided.

Assembly requires no knowledge of logic techniques or Karnaugh maps. If you can send Morse, you can program the matrix. Programming and assembly time is three or four hours. The memory kit

yields three programs. A typical contest set are:

5NN03 DE W1DTY
CQ TEST CQ TEST DE W1DTY
DE W1DTY

In addition to operating convenience, the new message generator is expected to raise cw contest scores by boosting operator efficiency. The EK-39MK Diode Memory Kit is available now from dealers or direct from the manufacturer and is priced at $49.95. (The EK-39M is $179.95.) For more information write Curtis Electro Devices, Box 4090, Mountain View, California 94040, or use *check-off* on page 94.

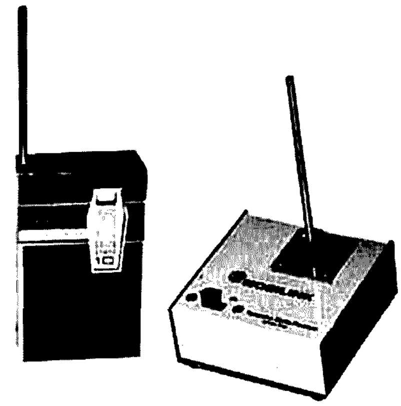
motorola semiconductor kits

Two new construction kits with special "how to" project brochures have been introduced by Motorola HEP through its nationwide network of semiconductor distributors. The HEK-3 Radio Amateur Hobby Kit, retailing for $5.95, contains two rf/i-f linear integrated circuits and an RTL integrated circuit in addition to an HMA-36 special project brochure. Circuits in the brochure include a 6-meter preamp, 6-meter a-m modulator, 40-meter transmitter, video amplifiers and a microphone amplifier.

The HEK-4 Home Handyman Hobby Kit, retailing for $4.95, contains a silicon-controlled rectifier, a unijunction transistor, a silicon rectifier, a photo transistor and a silicon npn transistor. Special project brochure HMA-37 that comes with this kit shows 11 applications including a light dimmer, electronic timer, burglar alarm and code-practice oscillator.

The project brochures are free at HEP semiconductor distributors. HEP is Motorola's sales program for making semi-conductor devices readily available to the experimenter through a nationwide network of authorized suppliers.

mobilink



The Mobilink is a unique new product that will have wide acceptance among two-way radio users in the commercial field, CB, amateurs, and other operators who require constant monitoring of their mobile receivers. The Mobilink is helpful when out of the car or out of hearing range while on business calls, coffee breaks or lunch.

Mobilink is a low powered a-m transmitter used with a pocket receiver as an accessory for mobile communications, base stations, paging and intercom extensions. All calls and the exact content of messages will be heard up to 1/4 mile away on the small receiver. This is a great improvement over the systems that blow the car horn, or turn on the car lights. Mobilink has a greater usable distance, will not disturb others and will not discharge the car battery.

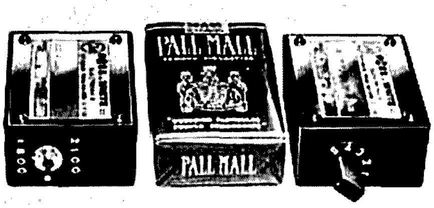
The Mobilink features easy installation – only connection needed is a pair of wires to the speaker; the internal 9-volt battery has shelf life in standard service due to the vox circuit that turns on the Mobilink transmitter only when a signal is received on the communications receiver. It can be used with all code, scramble and squelch systems without modification because it uses audio instead of an rf signal.

The Mobilink has 75 milliwatts input power, an antenna that adjusts from 5 to 18 inches, small size, weighs only 11 ounces, frequency of 27.263 MHz or any CB channel on request, crystal controlled, input impedance of 3 to 8 ohms, heavy aluminum construction, all solid-state

devices and made in U. S. A. Price is $44.95 complete with crystal, battery, connecting wire and antenna.

The Mobilink receiver is a small pocket size portable transistor circuit with 18-inch collapsible antenna, rf tuned circuit, crystal controlled, speaker, earphone, agc performance, 9-volt battery and carrying case. Comes complete ready to use with crystal for $24.95 postpaid. Mobilink transmitter and receiver set is $69.90 postpaid. Units and information are available from Herbert Salch & Co., Marketing Division of Tompkins Radio Products, Woodsboro, Texas 78393, or use *check-off* on page 94.

continuous-tone encoder



Automatically activated signal-tone encoders for use with two-way radio equipment have been announced by the Ross and White Company. Keyed by closing the transmitter microphone switch, the encoder generates a short duration tone burst which modulates the transmitter, and automatically activates tone-access repeaters, tone-operated receiver squelch circuits or other signalling devices.

The two models are the TE-2, with two switch-selected output tones, and the TE-5, with five tones. Frequency range is field adjustable from 1600 to 2800 Hz. The 0.5 second output tone burst is a 6-V p-p open-circuit sine wave, with output impedance adjustable in three steps. The all solid-state circuit uses commercial-

grade American-made components and silicon semi-conductors on an epoxy-fiberglass circuit board.

The encoders and their 9-V batteries are housed in 2-1/8 x 1-1/8 x 3-1/4-inch plastic cases, and are connected to the transmitter microphone circuit with a 15-inch shielded cable. Installation typically requires 20 minutes or less. An aluminum bracket is provided for mounting the encoder.

The price of the TE-2 is $29.95, and for the TE-5, $39.95. For further information, contact Ross and White Company, 50 West Dundee Road, Wheeling, Illinois 60090, or use *check-off* on page 94.

QRPP magazine

Ade Weiss, K8EEG, long time QRP enthusiast, has launched a new bi-monthly, now in its second year, for the QRPP amateur radio operator. To the uninitiated, QRPP is an extension of the Q-signal QRP which is applied to amateur operation of less than five watts. The new magazine, *The Milliwatt: National Journal of QRPP,* is devoted exclusively to under-five-watt amateur radio, and has featured technical articles, construction projects, operating news, QRPP WAS standings and QRPP log selections. The first volume includes construction projects for transistor transmitters and receivers for 3.5, 7 and 14 MHz, QRPP wattmeters, dummy loads and SWR meters, articles on propagation, operating procedures and other applications of general theory to the specific requirements of very low power operation. *The Milliwatt* is published six times a year; subscriptions are $3.00 annually, $3.40 first-class mail. A reprint of the entire Volume I is available, $4.00 postpaid. Subscriptions and general queries to Wes Mattox, K6EIL/2, 115 Park Avenue, Binghamton, New York 13903. Articles and operating news to Adrian Weiss, Editor, *The Milliwatt,* Meckling, South Dakota 57044.

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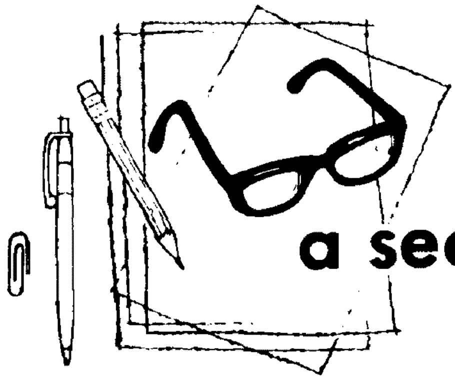

a second look

by jim fisk

new phone bands

If the **FCC's** newly **proposed** amateur radio frequency allocations are adopted, Advanced and Extra class amateurs can look forward to much wider phone bands. The new proposal, so new a number had not been assigned as of press time, provides for more operating space on all five high-frequency bands — 80, 40, 20, 15 and 10 meters. On 80 meters, for example, the Extra-class phone band would begin at 3750 kHz; the Advanced-class band would start at 3775 kHz while the General-class phone privileges would be increased to 3875 to 4000 kHz.

On the other high-frequency bands, the Extra-class phone segments would begin at 7150, 14150, 21200, and 28350 kHz. The Advanced-class phone segments would be at 3775, 7175, 14175, 21225 and 28375 kHz. Under the new plan General-class phone privileges would be expanded to start at 3875, 7225, 14250, 21325 and 28500 kHz. In addition to increased operating space on 80, 40, 20 and 15 meters, note the new Advanced and Extra class phone allocation on 10 meters.

Also included in the new FCC proposal is an all-new Advanced and Extra-class phone sub-band on 40 meters, from 7075 to 7100 kHz. This is specifically designed for radiotelephone communications between the United States and Europe and Africa. The new proposal would also reduce the width of the exclusive Extra-class CW segments on each of the high-frequency bands — from 25 to 10 kHz.

In addition, under the new plan Novices would gain an additional 100 kHz of CW operating space on 10 meters — between 28150 and 28250 kHz.

Remember, these phone bands are still in the proposal stage. For more details I suggest you listen to the W1AW broadcasts from A.R.R.L. They are able to disseminate the pertinent details much more rapidly than a monthly magazine. Next month we will publish the complete text of the FCC proposal in *ham radio*. In the meantime, consider all the ramifications of this plan, discuss it with your friends and at your local radio club, and submit your comments to the FCC before June 1, 1971 (remember that all comments must be submitted in 14 copies).

circuits and techniques

On page 34 of this issue you'll find the first installment of a new monthly column, **circuits and techniques**. This new addition to *ham radio,* written by Edward M. Noll, W3FQJ, will endeavor to keep you up to date on the latest developments in communications and electronics. In addition to state-of-the-art developments, Ed will cover little known circuits and experimental techniques that are useful to amateurs and experimenters. In many cases overseas amateurs devise solutions to electronic problems that receive little publicity in this country. Ed will be reviewing many of the foreign electronics publications in an effort to keep up to date on this phase of amateur radio.

Ed Noll is well qualified for this new assignment with *ham radio.* As a long-time radio amateur, consulting engineer and accomplished author of technical articles and books, Ed brings to *ham radio* a background of translating complex electronics ideas into easy-to-understand text. His latest book, "Solid-State QRP Projects," was recently published by Editors and Engineers.

Jim Fisk, W1DTY
editor

practical vhf and uhf coil-winding data

The practical coil-winding data presented here provides complete details on inductors from 2 nanohenries to 1 microhenry

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This article contains computer generated data for building inductors from 2 to 1000 nanohenries (1 nanohenry = .001 μH). Since no calculations are involved, it is a simple matter to scan the tables, and select the inductor that best meets your particular requirements. The first part of the article describes single-layer solenoids from 10 to 1000 nH; the last part describes straight-wire inductors above a chassis that range from 2 to 100 nH.

vhf inductors

Many vhf experimenters have developed a sixth sense for winding rf coils — they've had to, since there doesn't seem to be any convenient coil winding data for this part of the spectrum. (The ARRL Lightning Calculator stops at 1 μH and the Allied Coil Winding Calculator stops at 0.1 μH.)

The typical design procedure is to wrap some wire around a pencil (a coil form is also permitted) and trim the coil to resonance with the aid of a grid-dip meter and fixed capacitor. However, it takes a fair amount of experience to select the proper wire size and coil diameter that will give the desired inductance and still have reasonable Q and low capacitance.

Tables 1, 2, 3 and **4** describe coils of 1 to 10 turns wound with an inside diameter of 1/8 to ½ inch.* Because of their size, these coils are especially attractive for use with solid-state receivers and transmitters.

design philosophy

The goal is an inductor that has high Q, low capacitance and compact size. Low coil capacitance means the inductor will have a high self-resonant frequency, and therefore, a more useful frequency range. This can be achieved by a single-layer solenoid with adequate turns spacing. A good rule of thumb is to have a space equal to the wire diameter between adjacent turns with coil length about 1.5

*The tables were computed from the formula

$$L = \frac{\left(\frac{ND}{2}\right)^2}{4.5D + 10l} \qquad (1)$$

where L is inductance, D is coil diameter and l is coil length. This formula approximates the low-frequency inductance of a coil in free space. However, after building a few coils and measuring their inductance with a Boonton 250A RX meter at 100 MHz it appears that the error is only 10% for most coils.

times the coil diameter. The result is a coil with low capacitance and reasonable Q. All coils computed in the tables have turns spacing equal to the diameter of the wire used; as a check, the overall length of the coil is also given.

Those coils whose length is 1 to 2 times diameter are shown in bold type since they are considered to be optimum. By scanning the tables you can see that any inductance can be obtained by an optimum coil.

All calculated inductances were rounded off to the nearest 10 nanohenries. This means that the error of values below 30 nH will be ±5 nH. This seemed sufficient since adjacent objects will introduce errors into the free-space design anyway. Below 10 nH it is usually easier to build straight-wire inductors.

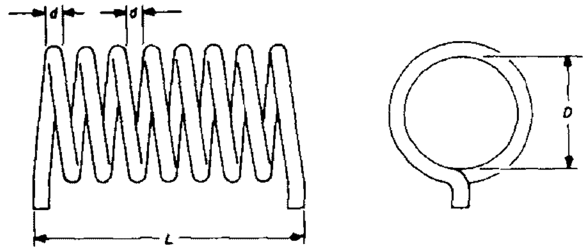


fig. 1. Airwound coil showing construction dimensions.

table 1. Coil data for 0.125-inch diameter airwound coils. Bold-face values represent optimum designs

wire size	1	2	3	4	5	6	7	8	9	10	
18	**5**	**10**	**20**	30	30	40	50	60	70	70	nH
	.12	**.20**	**.28**	.36	.44	.52	.60	.69	.77	.85	inch
20	5	**10**	**20**	30	40	50	50	60	70	80	nH
	.10	**.16**	**.22**	.29	.35	.42	.48	.54	.61	.67	inch
22	5	**10**	**20**	**30**	40	50	60	70	80	90	nH
	.08	**.13**	**.18**	**.23**	.28	.33	.38	.43	.48	.53	inch
24	5	10	**20**	**30**	**50**	**60**	70	80	100	110	nH
	.06	.10	**.14**	**.18**	**.22**	**.26**	.30	.34	.38	.42	inch

(number of turns: 1–10)

table 2. Coil data for 0.25-inch diameter airwound coils. Bold-face values represent optimum designs.

Wire size	1	2	3	4	5	6	7	8	9	10	
12	**10**	**20**	**30**	50	70	80	100	120	130	150	nH
	.24	**.40**	**.57**	.73	.89	1.05	1.21	1.37	1.54	1.7	inch
14	**10**	**20**	**40**	50	70	90	110	130	150	170	nH
	.19	**.32**	**.45**	.58	.71	.83	.96	1.09	1.22	1.35	inch
16	10	**20**	**40**	**60**	**80**	100	120	140	170	190	nH
	.15	**.25**	**.36**	**.46**	**.56**	.66	.76	.86	.97	1.07	inch
18	10	**30**	**50**	**70**	**90**	**120**	140	170	190	220	nH
	.12	**.20**	**.28**	**.36**	**.44**	**.52**	.60	.69	.77	.85	inch
20	10	30	**50**	**80**	**100**	**130**	**160**	**190**	220	250	nH
	.10	.16	**.22**	**.29**	**.35**	**.42**	**.48**	**.54**	.61	.67	inch
22	10	30	**60**	**90**	**120**	**150**	**180**	**220**	**250**	290	nH
	.08	.13	**.18**	**.23**	**.28**	**.33**	**.38**	**.43**	**.48**	.53	inch
24	10	30	**60**	**100**	**130**	**170**	**210**	**250**	**290**	**340**	nH
	.06	.10	**.14**	**.18**	**.22**	**.26**	**.30**	**.34**	**.38**	**.42**	inch

(number of turns: 1–10)

table 3. Coil data for 0.375-inch diameter airwound coils. Bold-face values represent optimum designs.

wire size	number of turns 1	2	3	4	5	6	7	8	9	10	
10	10	30	60	80	110	130	160	190	210	240	nH
	.31	.51	.71	.92	1.12	1.32	1.53	1.73	1.94	2.14	inch
12	10	30	60	90	120	150	180	210	240	270	nH
	.24	.40	.57	.73	.89	1.05	1.21	1.37	1.54	1.70	inch
14	10	40	70	100	130	170	200	240	280	310	nH
	.19	.32	.45	.58	.71	.83	.96	1.09	1.22	1.35	inch
16	10	40	70	110	150	190	230	270	320	360	nH
	.15	.25	.36	.46	.56	.66	.76	.86	.97	1.07	inch
18	10	40	80	130	170	220	270	320	370	420	nH
	.12	.20	.28	.36	.44	.52	.60	.69	.77	.85	inch
20	10	50	90	140	190	250	310	360	420	480	nH
	.10	.16	.22	.29	.35	.42	.48	.54	.61	.67	inch
22	20	50	100	160	220	280	350	420	490	560	nH
	.08	.13	.18	.23	.28	.33	.38	.43	.48	.53	inch
24	20	60	110	170	240	320	400	480	560	650	nH
	.06	.10	.14	.18	.22	.26	.30	.34	.38	.42	inch

table 4. Coil data for 0.5-inch diameter airwound coils. Bold-face values represent optimum designs.

wire size	number of turns 1	2	3	4	5	6	7	8	9	10	
10	20	50	80	120	160	200	250	290	330	380	nH
	.31	.51	.71	.92	1.12	1.32	1.53	1.73	1.93	2.14	inch
12	20	50	90	140	180	230	280	330	380	430	nH
	.24	.40	.57	.73	.89	1.05	1.21	1.37	1.54	1.70	inch
14	20	60	100	150	210	260	320	380	440	500	nH
	.19	.32	.45	.58	.71	.83	.96	1.09	1.22	1.35	inch
16	20	60	110	170	240	300	370	440	510	580	nH
	.15	.25	.36	.46	.56	.66	.76	.86	.97	1.07	inch
18	20	70	130	190	270	340	420	500	590	670	nH
	.12	.20	.28	.36	.44	.52	.60	.69	.77	.85	inch
20	20	70	140	210	300	390	480	580	680	780	nH
	.10	.16	.22	.29	.35	.42	.48	.54	.61	.67	inch
22	20	80	150	240	340	440	550	660	780	900	nH
	.08	.13	.18	.23	.28	.33	.38	.43	.48	.53	inch
24	20	80	160	260	370	490	620	750	890	1030	nH
	.06	.10	.14	.18	.22	.26	.30	.34	.38	.42	inch

example 2. A 50-nH coil is required for a 20 watt transmitter.

possibility		comment
0.125 diam	5 turns no. 24	poor choice at this power level
0.250 diam	4 turns no. 12 or no. 14	fair choice — only slightly out of optimum region
0.250 diam	3½ turns no. 16	marginal at this power level
0.250 diam	3 turns no. 18	marginal at this power level
0.375 diam	2.7* turns no. 10	good choice
0.375 diam	2.7 turns no. 12	good choice
0.375 diam	2.3 turns no. 14	good choice
0.500 diam	2 turns no. 10	good choice
0.500 diam	2 turns no. 12	good choice

example 3. Same 50-nH coil as in example 2 but this time it is required for a receiver

0.125 dia.	5 turns no. 24	good choice, compact size
0.250 dia.	3.5 turns no. 16	good choice
0.250 dia.	3 turns no. 18	good choice
0.375 dia.	2.7 turns no. 10	good choice, but large size may add too much capacitance to circuit

using the tables

The tables are intended for air-core coils whose dimensions are indicated in fig. 1. Each table describes coils wound with a different inside diameter. Wire size and number of turns are specified along the edge of the chart. The data within the table is inductance in nanohenries (on top) and coil length in inches (below). The use of the inductance tables is best illustrated by several practical examples.

example 1. What is the inductance of 5 turns no. 18 wire, 0.25-inch diameter, wound with spacing equal to wire diameter? From **table 2,** opposite no. 18, and below 5 turns, you find this coil has 90-nH inductance and is 0.44 inches long.

A coil of given inductance can be easily designed by scanning the optimum regions (bold-faced type) of each table. If the exact value is not found, the inductance may be mentally interpolated by changing the turns by a fraction or by compressing or expanding coil length.

uhf inductors

As you can see from **tables 1, 2, 3** and 4, it is impractical to wind coils less than 10 nH. For less than 10 nH the inductance of a straight piece of wire is sufficient. Quarter-wavelength resonators are common in microwave work and may be considered as an inductance in parallel with distributed capacitance.

Full-sized ¼-wave resonators are useful above 1 or 2 GHz because of their convenient size and high Q. But at 432 MHz or even 1296, the designer may want a more compact resonator. This can be accomplished by shortening the length needed for ¼-wave resonance and making up for the decreased inductance by adding external capacitance.

Obviously this is a design trade-off resulting in a lower Q, since $Q = X_L/R$, and decreased inductance means lowered Q. However, you have gained more compact size: e. g., 432-MHz tank circuits may be built 1 or 2 inches long as compared with a full quarter-wavelength of 7 inches. You have also avoided an impedance-matching problem since connecting circuitry will usually be capacitive anyway. In a transistor tank circuit the collector capacitance, tuning capacitor and coil capacitance are combined. Output is taken by either capacitor-divider coupling, transformer coupling or tapping

*Instead of winding fractional turns, the coil may be wound with 3 turns and "stretched" to the desired inductance.

down on the coil. (Motorola has an excellent application note for rf transistor design.)[2]

Tables 5, 6 and **7** contain computed data describing a wire of diameter D and length L, spaced height H above a ground plane as shown in **fig. 2.*** Wire size, height above ground and length in inches are specified along the edge of the inductance tables. The data within the table is inductance (nH) on top, capacitance (pF) in the middle, self-resonance (GHz) on

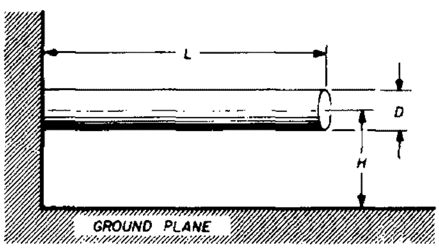


fig. 2. Length of wire above a ground plane exhibits both inductance and capacitance. Dimensions are used to calculate values.

the bottom. As before, the use of these tables is best illustrated by several typical examples.

example 4. What are the characteristics of a 2-inch length of no. 10 wire, spaced 0.25 inch above a ground plane? From **table 5**, a 2-inch length of no. 10 wire has 21-nH inductance in parallel with 0.7 pF. Self-resonant frequency 1.2 GHz (1200 MHz).

design philosophy

A quick scan of tables **5, 6** and **7** reveals some interesting phenomena that should be kept in mind when laying out circuits. For example, moving the inductor closer to a ground plane increases its capacitance. Not so obvious is the fact that this also decreases inductance. The inductor and the ground plane may be considered to be a transformer with a shorted secondary. Hence, increased coupling results in less inductance. It turns out that the capacitance changes more than inductance, and the net result is a lower resonant frequency.

Moving the inductor away from the chassis will raise the Q. Beyond a height of one inch, however, the computed L and C rapidly approaches the free-space inductance as a limit, and the law of diminishing returns applies.

Considering the resonator as a transmission line, its characteristic impedance is $Z_O = \sqrt{L/C}$. Thus, moving the ¼-wave resonator too far from the chassis will raise its impedance to match the approximately 377-ohm radiation resistance of space. Then the resonator will then behave more like an antenna than a resonator.

Adding additional ground planes at right angles to form a coaxial cavity around the wire lowers the resonant frequency by about 10%. This implies that L and C have changed by more than that amount since they move in opposite directions. An estimate of the inductance and capacitance of a coaxial shielded wire can be made by considering it simply as a wire that is closer to a single ground plane.

*The inductance values shown in **tables 5, 6** and **7** were calculated from the formula

$$L = .0116967\,(\log 4H/D + \log \frac{A}{B}) + .00508\,(B - A + \mu 1 \delta - 2H + \frac{D}{2})$$

$$\text{where } A = 1 + \sqrt{1^2 + \frac{D^2}{4}}$$

$$B = 1 + \sqrt{1^2 + 4H^2}$$

$$\mu \text{ (permeability)} = 1$$

Skin effect, because of its very small value, was neglected. The capacitance of the straight wire above a ground plane was calculated from

$$C = \frac{\pi \epsilon l}{\ln \left(\frac{4H}{D} - 1\right)}$$

where ϵ is permitivity. As a check, capacitance measurements were made on a Boonton 250A RX meter operating at 1 MHz. Readings were within 0.4 pF of the calculated values.

Next, the circuit of **fig. 2** was duplicated, and a signal generator and rf detector were loosely coupled to the resonator. For each case measured the self-resonant frequency was within 20% of that calculated from the computed inductance and capacitance. It is also gratifying that there is some correlation between the computed LC resonant frequency and resonance of ¼-wave transmission lines.

table 5. Inductance of wire 0.25 inch above a ground plane. Upper value is inductance in nH, middle value is capacitance in pF, lower value is self-resonant frequency in GHz.

wire	length (inches)										
size	0.5	1.0	1.5	2.0	2.5	3.0	3.5	4.0	4.5	5.0	
	2	5	9	12	15	19	22	26	29	33	nH
2	.4	.7	1.1	1.5	1.8	2.2	2.5	2.9	3.3	3.6	pF
	5.3	2.4	1.5	1.1	.9	.7	.6	.5	.5	.4	GHz
	3	6	10	14	18	22	26	30	34	38	nH
4	.3	.6	.8	1.1	1.4	1.7	2	2.3	2.5	2.8	pF
	5.4	2.4	1.6	1.2	.9	.8	.6	.6	.5	.4	GHz
	3	7	12	17	21	26	30	35	40	44	nH
6	.2	.5	.7	.9	1.2	1.4	1.6	1.9	2.1	2.3	pF
	5.5	2.5	1.6	1.2	.9	.8	.7	.6	.5	.5	GHz
	4	9	14	19	24	29	34	40	45	50	nH
8	.2	.4	.6	.8	1	1.2	1.4	1.6	1.8	2	pF
	5.5	2.5	1.6	1.2	.9	.8	.7	.6	.5	.5	GHz
	4	10	15	21	27	33	38	44	50	56	nH
10	.2	.4	.5	.7	.9	1.1	1.2	1.4	1.6	1.8	pF
	5.4	2.5	1.6	1.2	1	.8	.7	.6	.5	.5	GHz
	5	11	17	23	30	36	42	49	55	62	nH
12	.2	.3	.5	.6	.8	.9	1.1	1.3	1.4	1.6	pF
	5.4	2.5	1.6	1.2	1	.8	.7	.6	.5	.5	GHz
	5	12	19	26	33	40	47	54	61	67	nH
14	.1	.3	.4	.6	.7	.9	1	1.1	1.3	1.4	pF
	5.4	2.5	1.6	1.2	1	.8	.7	.6	.5	.5	GHz
	6	13	21	28	36	43	51	58	66	73	nH
16	.1	.3	.4	.5	.7	.8	.9	1	1.2	1.3	pF
	5.3	2.5	1.6	1.2	1	.8	.7	.6	.5	.5	GHz
	6	14	22	30	38	47	55	63	71	79	nH
18	.1	.2	.4	.5	.6	.7	.8	1	1.1	1.2	pF
	5.3	2.5	1.6	1.2	1	.8	.7	.6	.5	.5	GHz
	7	15	24	33	41	50	59	68	76	85	nH
20	.1	.2	.3	.5	.6	.7	.8	.9	1	1.1	pF
	5.3	2.5	1.6	1.2	1	.8	.7	.6	.5	.5	GHz
	7	17	26	35	44	54	63	72	82	91	nH
22	.1	.2	.3	.4	.5	.6	.7	.8	.9	1.1	pF
	5.2	2.5	1.6	1.2	1	.8	.7	.6	.5	.5	GHz
	8	18	27	37	47	57	67	77	87	97	nH
24	.1	.2	.3	.4	.5	.6	.7	.8	.9	1	pF
	5.2	2.5	1.6	1.2	1	.8	.7	.6	.5	.5	GHz

Uhf resonators are usually made from the larger diameter wires, but data for wires smaller than no. 18 is included mainly for estimating component-lead inductance. The resonant frequency given in the tables sets the upper limit at which the inductor may be used; above resonance it acts like a capacitor. The inductor should be chosen so that with the added external circuit capacitance the LC combination will resonate at the desired frequency.

example 5. It is desired to design a transistor tank circuit for 430 MHz as shown in **fig. 3.** The transistor has an output capacitance of 3 pF, and

table 6. Inductance of wire 0.5 inch above a ground plane. Upper value is inductance in nH, middle value is capacitance in pF, lower value is self-resonant frequency in GHz.

wire size	0.5	1.0	1.5	2.0	2.5	3.0	3.5	4.0	4.5	5.0	
						length (inches)					
	3	7	12	17	22	27	32	38	43	48	nH
2	.2	.4	.6	.8	1	1.2	1.4	1.6	1.8	2	pF
	6.3	2.7	1.7	1.3	1	.8	.7	.6	.5	.5	GHz
	3	8	14	19	25	31	36	42	48	54	nH
4	.2	.4	.5	.7	.9	1.1	1.2	1.4	1.6	1.8	pF
	6.2	2.7	1.7	1.3	1	.8	.7	.6	.5	.5	GHz
	4	9	15	21	28	34	40	47	53	59	nH
6	.2	.3	.5	.6	.8	.9	1.1	1.3	1.4	1.6	pF
	6.1	2.7	1.7	1.3	1	.8	.7	.6	.5	.5	GHz
	4	10	17	24	31	37	44	51	58	65	nH
8	.1	.3	.4	.6	.7	.9	1	1.1	1.3	1.4	pF
	6	2.7	1.7	1.3	1	.8	.7	.6	.5	.5	GHz
	5	11	19	26	33	41	48	56	64	71	nH
10	.1	.3	.4	.5	.7	.8	.9	1.1	1.2	1.3	pF
	5.9	2.7	1.7	1.3	1	.8	.7	.6	.5	.5	GHz
	5	13	20	28	36	44	53	61	69	77	nH
12	.1	.2	.4	.5	.6	.7	.8	1	1.1	1.2	pF
	5.8	2.7	1.7	1.2	1	.8	.7	.6	.5	.5	GHz
	6	14	22	31	39	48	57	65	74	83	nH
14	.1	.2	.3	.5	.6	.7	.8	.9	1	1.1	pF
	5.8	2.6	1.7	1.2	1	.8	.7	.6	.5	.5	GHz
	6	15	24	33	42	51	61	70	79	89	nH
16	.1	.2	.3	.4	.5	.6	.7	.8	.9	1.1	pF
	5.7	2.6	1.7	1.2	1	.8	.7	.6	.5	.5	GHz
	7	16	26	35	45	55	65	75	85	94	nH
18	.1	.2	.3	.4	.5	.6	.7	.8	.9	1	pF
	5.6	2.6	1.7	1.2	1	.8	.7	.6	.5	.5	GHz
	7	17	27	38	48	58	69	79	90	100	nH
20	.1	.2	.3	.4	.5	.6	.7	.7	.8	.9	pF
	5.6	2.6	1.7	1.2	1	.8	.7	.6	.5	.5	GHz
	8	18	29	40	51	62	73	84	95	106	nH
22	.1	.2	.3	.4	.4	.5	.6	.7	.8	.9	pF
	5.5	2.6	1.7	1.2	1	.8	.7	.6	.5	.5	GHz
	9	19	31	42	54	65	77	89	100	112	nH
24	.1	.2	.3	.3	.4	.5	.6	.7	.8	.8	pF
	5.5	2.6	1.7	1.2	1	.8	.7	.6	.5	.5	GHz

the two impedance-matching variable, capacitors are assumed to present an average capacitance of 4 pF at the collector. Thus, total capacitance will be 7 pF plus inductor capacitance. An LC nomograph (fig. 4) indicates that 20 nH will resonate with 7 pF at 425 MHz.

The date for no. 14 wire spaced 0.25 inch above a ground plane (table 5) shows that a 1½-inch length has 17-nH inductance and 10.5 pF capacitance. Therefore, the tank circuit consists of 19 nH in parallel with 17.4 pF and has a midrange resonance of 424 MHz.

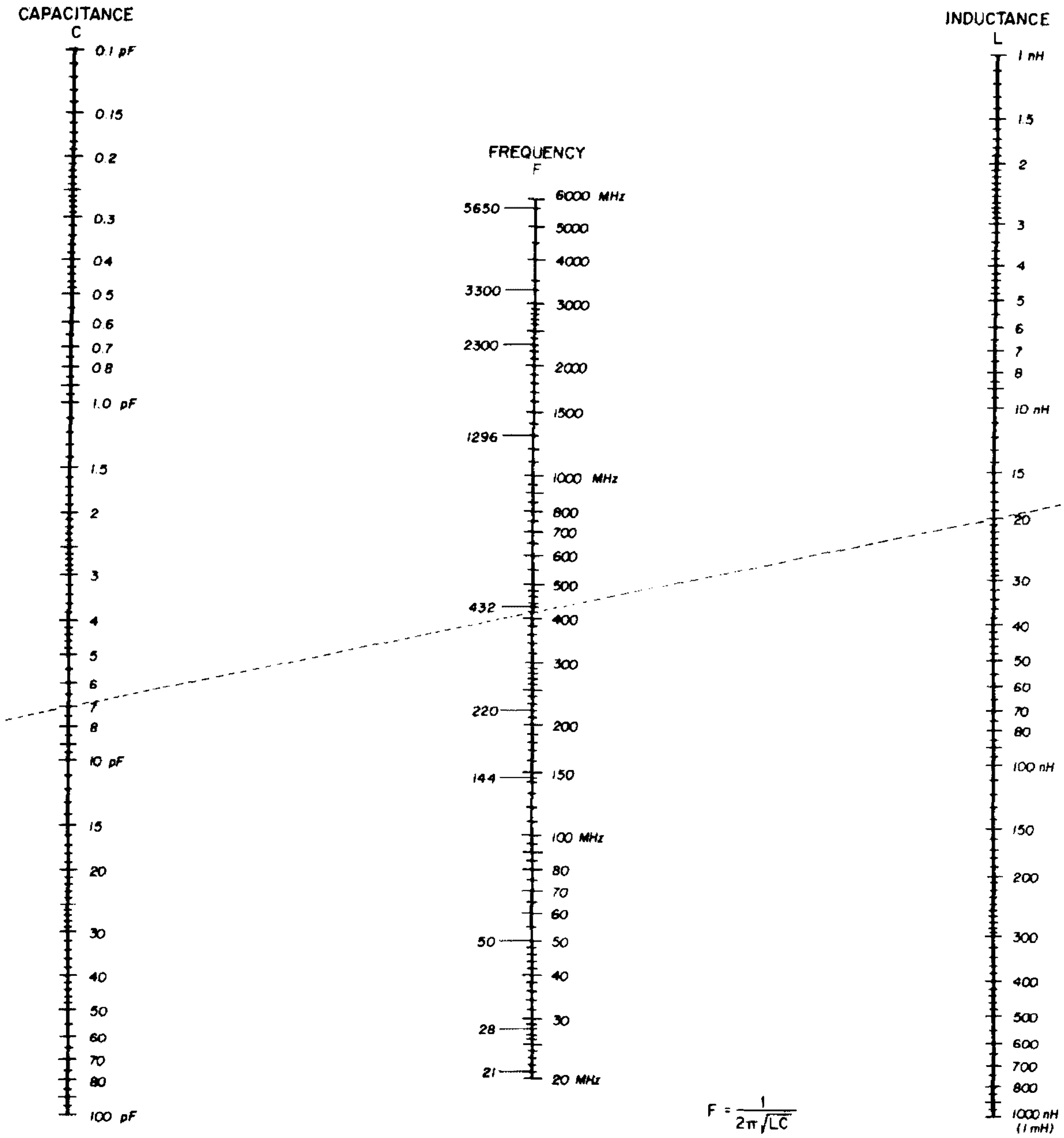
table 7. Inductance of wire 1.0 inch above a ground plane. Upper value is inductance in nH, middle value is capacitance in pF, lower value is self-resonant frequency in GHz.

wire size	0.5	1.0	1.5	2.0	2.5	3.0	3.5	4.0	4.5	5.0	
	length (inches)										
	3	8	14	21	27	34	41	47	54	61	nH
2	.1	.3	.4	.6	.7	.9	1	1.1	1.3	1.4	pF
	7.1	3	1.9	1.3	1	.9	.7	.6	.6	.5	GHz
	3	9	16	23	30	37	45	52	59	67	nH
4	.1	.3	.4	.5	.7	.8	.9	1.1	1.2	1.3	pF
	6.9	3	1.8	1.3	1	.9	.7	.6	.6	.5	GHz
	4	10	18	25	33	41	49	57	65	73	nH
6	.1	.2	.4	.5	.6	.7	.8	1	1.1	1.2	pF
	6.7	2.9	1.8	1.3	1	.8	.7	.6	.6	.5	GHz
	4	11	19	27	36	44	53	61	70	78	nH
8	.1	.2	.3	.5	.6	.7	.8	.9	1	1.1	pF
	6.5	2.9	1.8	1.3	1	.8	.7	.6	.5	.5	GHz
	5	13	21	30	39	48	57	66	75	84	nH
10	.1	.2	.3	.4	.5	.6	.7	.8	.9	1.1	pF
	6.4	2.8	1.8	1.3	1	.8	.7	.6	.5	.5	GHz
	5	14	23	32	41	51	61	71	80	90	nH
12	.1	.2	.3	.4	.5	.6	.7	.8	.9	1	pF
	6.3	2.8	1.8	1.3	1	.8	.7	.6	.5	.5	GHz
	6	15	24	34	44	55	65	75	86	96	nH
14	.1	.2	.3	.4	.5	.6	.7	.7	.8	.9	pF
	6.2	2.8	1.8	1.3	1	.8	.7	.6	.5	.5	GHz
	7	16	26	37	47	58	69	80	91	102	nH
16	.1	.2	.3	.4	.4	.5	.6	.7	.8	.9	pF
	6.1	2.8	1.8	1.3	1	.8	.7	.6	.5	.5	GHz
	7	17	28	39	50	62	73	85	96	108	nH
18	.1	.2	.3	.3	.4	.5	.6	.7	.8	.8	pF
	6	2.7	1.7	1.3	1	.8	.7	.6	.5	.5	GHz
	8	18	30	41	53	65	77	89	101	114	nH
20	.1	.2	.2	.3	.4	.5	.6	.6	.7	.8	pF
	5.9	2.7	1.7	1.3	1	.8	.7	.6	.5	.5	GHz
	8	19	31	44	56	69	81	94	107	119	nH
22	.1	.2	.2	.3	.4	.5	.5	.6	.7	.8	pF
	5.8	2.7	1.7	1.3	1	.8	.7	.6	.5	.5	GHz
	9	21	33	46	59	72	85	99	112	125	nH
24	.1	.1	.2	.3	.4	.4	.5	.6	.7	.7	pF
	5.8	2.7	1.7	1.3	1	.8	.7	.6	.5	.5	GHz

summary

It is one thing to design on paper but uhf and microwave work always require a certain amount of "cut and try." The approximations made and factors ignored in this article would probably send chills up the spine of a physicist. However, physicists don't have to design equipment and make things work.

Each piece of equipment is a unique problem. Armed with basic data and some mental fudge factors the designer can obtain a quick solution of reasonable



$$F = \frac{1}{2\pi\sqrt{LC}}$$

fig. 4. Resonant-frequency nomograph may be used to determine capacitor and inductor values over the range from 20 to 6000 MHz. The example indicates that 20 nH will resonate at 425 MHz with a 7-pF capacitor.

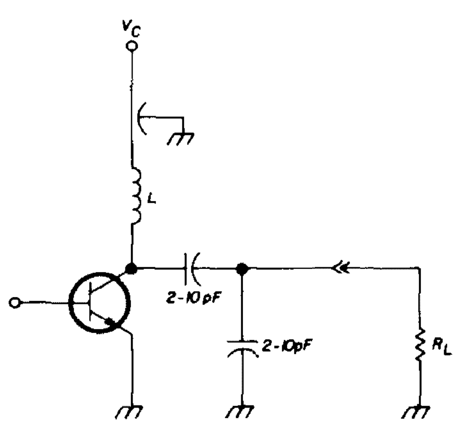


fig. 3. Typical 425-MHz tank circuit. Effective circuit capacitance of 7.4 pF will resonate with 19 nH at 424 MHz.

accuracy. Compared to that, an exact calculation is usually impractical.

references

1. "Reference Data for Radio Engineers," Fifth edition, ITT.
2. Frank Davis, "Matching Network Designs with Computer Solutions," Motorola Application Note AN-267, Motorola Semiconductor Products, Inc., Box 995, Phoenix, Arizona 85001.
3. Keith Henney, Radio Engineering Handbook, McGraw-Hill, New York.
4. John Ryder, "Network Lines and Fields," 2nd edition, 1949, Prentice-Hall, New York.

ham radio

how to use ferrite and powdered-iron for inductors

Hank Olson, W6GXN, Post Office Box 339, Menlo Park, California 94025

A complete description of ferromagnetic rods, toroid cores, pot cores, and broadband transformers, and how to use them in your circuits

"Just another hunk of Detroit Iron," says the sports-car buff as he is passed by a new Camaro. Detroit does use a good deal of iron in the manufacture of automobiles, but glass, plastics and nonferrous metals are replacing it in increasing quantities.

In electronics too, the drive is on to eliminate the iron and get everything down to one 100 x 100-mil silicon chip. Countless articles have been written on how to eliminate inductors (and their iron) from resonant circuits. Because of these efforts there have been some notable successes in replacing inductors with capacitors by means of feedback techniques.

I praise these new techniques, not as replacements for iron, but as welcome *additions* to the engineering bag of tricks. With the new all-silicon techniques, we can now come up with solutions to problems that were once the job for large iron components. Furthermore, these new techniques solve problems that iron could never be expected to touch. If you adopt this view you gain engineering flexibility and can solve any given problem the *best* way (although the use of iron may be the best way).

A brief look at conventional transformers and chokes will help you understand some of the basics. Any experimenter who has taken a transformer or choke apart (either for curiosity or rewinding purposes) is aware that such devices are made of laminated steel sections. These laminations are generally in two forms, some resembling the letter E and some the letter I. This is called the E-I core. The careful disassembler probably also noted that the E and I sections in a swinging choke or transformer were stacked as shown in **fig. 1A;** smoothing-choke E and I sections use the arrange-

ment of fig. 1B with the addition of a small piece of cardboard. If you ever tried reassembling the transformer, you probably found that it ran hotter than the original, even when no more power was being transformed.

magnetic circuits

To understand some of these observations, it might be a good idea to review a few basics of magnetic circuits. There are a number of ferro-magnetic substances, including iron, nickel and cobalt. These materials have a peculiar characteristic, that of high μ.

The concept of μ (mu) can be explained qualitatively by a simple experiment. Wind a coil of 50 turns of number-24 enameled wire on a cardboard mailing tube and measure its inductance. It should come out in the neighborhood of 80 μH. Now insert a ferrite rod from an old transistor radio antenna into the coil; you should see the inductance increase dramatically. While this is not a rigid definition of μ, it will give you a feel of its effect.

Another useful magnetic concept is that of reluctance. If you can envision the presence of the ferrite rod in our experimental coil as providing a low-impedance path for the magnetic field lines that flow axially inside the coil, reluctance *is* that impedance. Since magnetic field lines always close on themselves (flowing from north pole to south pole inside the coil and back again on the outside) it would seem that we could further reduce reluctance by continuing our ferrite rod around the outside of the coil as in fig. 2B, thereby increasing inductance. In fact, this is true, and the result *is* a coil with more inductance.

The E-I construction is simply an extension of the technique of fig. 2B where *two* low-reluctance return paths are used in parallel outside the coil as in fig. 2C. A modern version of fig. 2B can be found in tape-wound C core or toroid used in 400-Hz power transformers and dc-to-dc converters.

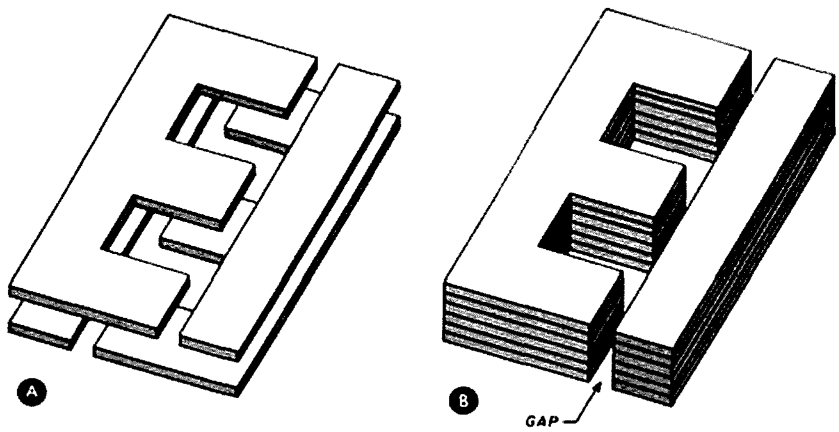


fig. 1. E-I transformer cores. Drawing in A shows how laminations are interleaved for minimum reluctance. B shows stacking of smoothing-choke laminations for insertion of cardboard air gap between E- and I-sections.

The mystery of why the rewound transformer ran hot can be understood in terms of eddy current loss. If the iron core was made of one *solid* piece of steel eddy-current loss would be very high. This occurs because the steel is an electrical conductor, and appears as a shorted turn in the same plane as the turns of the transformer winding. Such shorted turns are broken up by insulating layers between a laminated core — usually with some sort of varnish. In disassembly and reassembly of an E-I transformer core, the varnish is partially removed. Unless the reason for the varnish is understood, and it is replaced, the rewound transformer will run hot.

As transformers were called on to handle higher and higher frequencies it was found that the thickness of the laminations had to be made smaller. The reason for this is easily appreciated if we consider that the inductive reactance of a

shorted turn goes up with frequency. Thus, at higher frequencies a small diameter shorted turn can give the same inductive reactance as a large diameter shorted turn at lower frequencies.

The tape-wound cores that are available now have lamination thickness down to ¼ mil, and operate efficiently up to several thousand cycles per second. Of course, the thickness of the inter-lamination insulation must be decreased at the same rate as the laminate. Otherwise the amount of magnetic material per unit volume in the core will be decreased. Such a decrease in magnetic material per unit volume would lower the μ of the core.

Since the early days of telephone another type of magnetic core has been in use – the powdered-iron toroid. This form was suggested before 1900 by Oliver Heavyside as a solution to eddy-current loss.[1] By coating iron filings with insulation and compressing them into a doughnut shape, a non-conductive, closed-magnetic structure is created that is useful to fairly high frequencies. However, this technique also has its limitations because the size of the iron particles must decrease at higher frequencies, as must the thickness of the insulating coating.

ferrite

Ferrite is a relatively new material to be used in magnetic cores. Since ferrites are not pure metals like iron, nickel or cobalt they do not have high electrical conductivity. In fact, ferrites are made mostly of iron oxide plus some additive such as manganese; they are sintered at high temperature and become ceramic-like – hard and difficult to modify after formation. Because of their low bulk conductivity, they offer nearly ideal high-frequency core characteristics: low eddy-current loss *and* high μ.

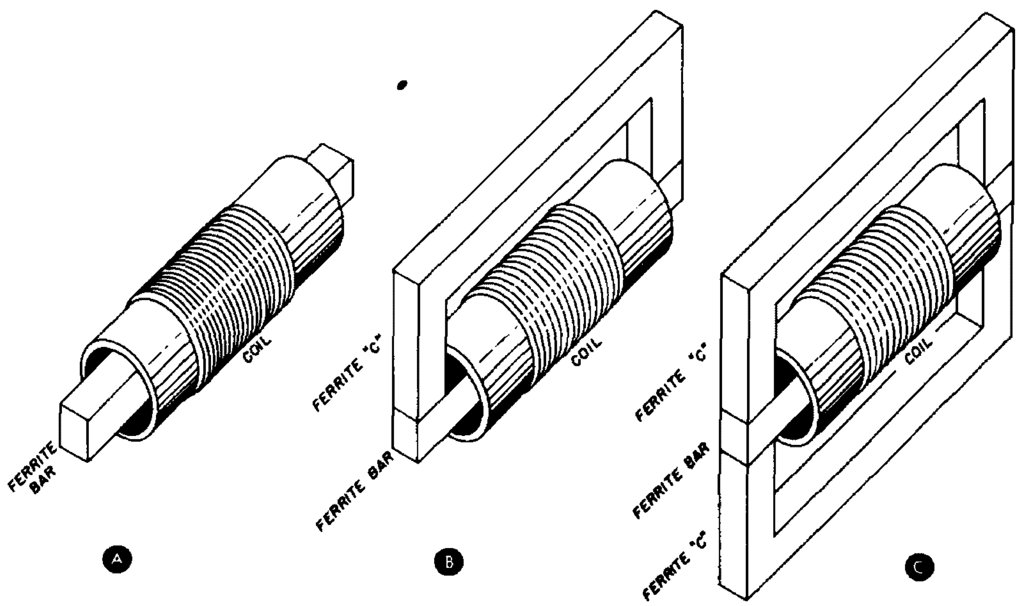


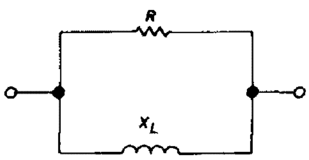
fig. 2. Inductance of simple coil is increased by ferrite core in A. By reducing the reluctance path as shown in B and C, inductance is further increased.

ferrite beads

The prospect of winding thousands of turns of fine wire on toroids, varnishing and restacking E-I sections, and doing other disagreeable tasks has probably discouraged most amateurs from ever building their own magnetic components. However, at audio frequencies and higher the situation is completely changed with modern magnetic cores. A useful ferrite-core choke can be simplified to a single turn! This, of course, reduces winding to a trivial business. The example I refer to is the ferrite bead.

Ferrite beads are made by several companies, but the beads most available

table 1. Equivalent parallel resistance and inductive reactance of Amidon Associates ferrite shielding beads (*measured* on Boonton 250A RX meter).



frequency (MHz)	resistance (ohms)	inductive reactance (ohms)
29	26.5	j43
50	29.7	j61
100	32.3	j80
144	33.4	j82
220	35.7	j72

to hams are sold in packets by Amidon Associates. The Amidon beads are useful for a host of choking applications; you simply slip a bead over a wire, and the wire becomes an rf choke. The nice thing about this rf choke is that it has very low Q, and is unlikely to create any nasty parasitic resonant circuits.

As an example, consider **fig. 3** which represents a B+ decoupling system for two stages of an rf unit. The two 1000-pF feedthrough capacitors with one inch of number-20 bare hookup wire between them form a parasitic resonance at about 45 MHz, and assure that adjacent stages are well coupled at that frequency. In a case where oscillation was rampant the simple expedient of slipping a ferrite bead over the wire cured the problem.

What strange properties do these little beads have to effect such a cure? You find the answer when you measure one on a wire; the results are shown in table **1**. Other beads (such as those made by Ferroxcube, Ferronics and Stackpole)

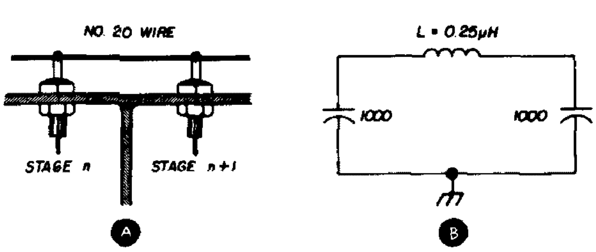


fig. 3. Typical B+ decoupling system is resonant at 45 MHz. Equivalent circuit is shown in B.

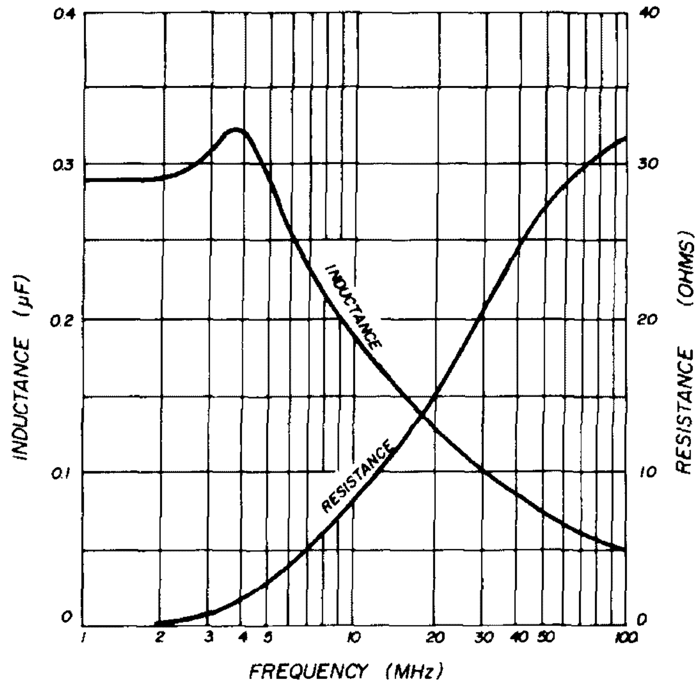


fig. 5. Typical values of equivalent series resistance and inductance for Ferronics ferrite beads.

behave similarly but they are not readily available except in large production quantities.

The fact that a typical bead and wire looks like a parallel combination of

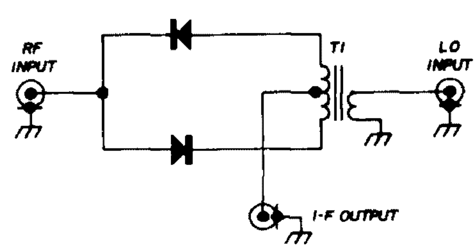


fig. 4. Simple balanced mixer using a ferrite-bead core. T1 is three turns no. 32 wire, trifilar wound on Amidon Associates ferrite bead. Two windings are cross connected to form center-tapped section. Diodes are Hewlett-Packard 5082-2800 or Motorola MBD101 (preferably matched).

35-ohms resistance and 75-ohms inductive reactance (35 + j75) is reminiscent of the 47-ohm resistor with a small coil wound on it for a plate or grid parasitic choke. Substituting a ferrite bead for a parasitic choke is just as satisfactory in practice as it appears to be in theory.

Beads are immensely useful for interstage decoupling at hf through uhf,

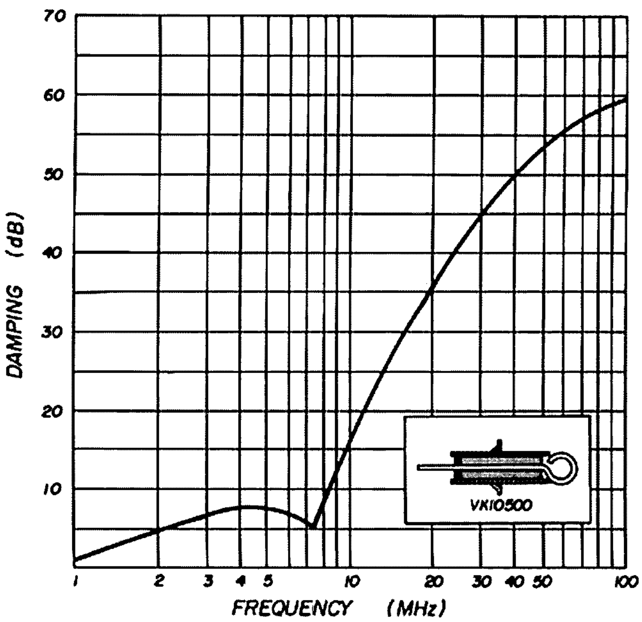


fig. 6. Performance of a Ferroxcube VK10500 decoupling filter when measured with a load impedance of 50 ohms.

powercord line decoupling, suppressing rf oscillations in audio and dc circuitry, and even in logic circuitry for decoupling splinters and spikes that cause logic blocks to communicate with each other. I've even used a ferrite bead as a tiny pulse transformer to create *very* short pulses in a blocking oscillator.[2,3] Fig. 4 shows a simple balanced mixer for vhf using a ferrite-bead core. There may be *better* cores for balanced (or doubly-balanced) mixers but none as inexpensive.[4]

More complex forms of shielding beads are also available as commercial items. Longer cylinders of ferrite are available from Ferroxcube, Stackpole, Siemens and Allen Bradley, with various numbers of axial holes in them. For instance, a Ferroxcube 56-390-31/4B two-hole bead could be used to decouple two filament leads.

You can insert a string of beads on a wire into a metal tube or section of copper braid for a lossy coaxial line that can be used for a really broadband rf rejection filter.[5] Different companies have different ways of specifying the performance of ferrite beads, but most of these specifications are of little direct use. The Ferronics beads are described by a simple graph showing the equivalent series R and L versus frequency of one bead (21-030-F) on a piece of number-20 wire. This is shown in fig. 5. Note that L decreases above 4 MHz, but since inductive reactance is proportional to frequency, it is held more or less constant.

A number of companies offer a combination ferrite bead and feedthrough capacitor for decoupling purposes. This low-pass L-network is very compact and effective if used properly. A curve of the Ferroxcube VK10500 is shown in fig. 6. The only precaution when using these devices is to get them in the right end-for-end configuration. Fig. 7 shows the correct and incorrect ways of using them. This is so the filter is in the LC, LC sequence required for a low-pass filter. When used incorrectly the two capacitances are connected with a piece of wire between them that results in a parasitic resonant circuit just as in fig. 3.

Many other types of ferrite cores are available besides beads. Toroids, pot cores, E-I combinations, U-I combinations, rods, binocular cores and many others produced by U. S., European and Japanese firms.

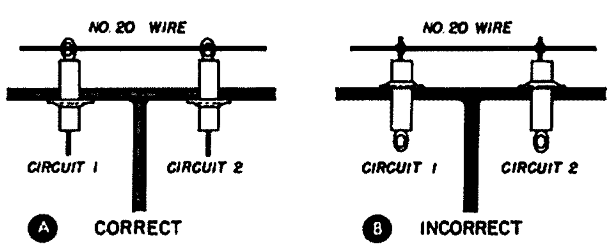


fig. 7. Correct use of Ferroxcube VK10500 decoupling filter is shown in A. Incorrect installation in B results in parasitic resonant circuit as in fig. 3.

toroid cores

The toroid is perhaps most familiar to hams.[6] However, it has several limitations; it is hard to wind (you need a bobbin or special toroid winding machine), and it is only available as a *closed* magnetic structure. The closed magnetic structure gives minimum reluctance, and hence the largest inductance per turn-

squared, but is easily saturated. The closed magnetic structure is similar to the *swinging* choke (which has no air gap) and E-I transformer. To avoid saturation, toroids are used in applications where very little dc current flows through the coils wound on them. This means that you'll find wide use of toroid inductors in audio preamps (in low- or band-pass filters), in receiver tuned circuits and in low power transmitter stages.

The really nice thing about toroids is their inherent magnetic shielding, since they have a closed magnetic structure. The entire magnetic field is virtually confined inside the toroid core *if you stay well below saturation.* This means that multistage filters and amplifiers can be built with toroid inductors without much concern for coupling between them (with the resultant instability). However, capacitive coupling between toroid coils must still be considered. This extreme magnetic field confinement is most apparent when you try to couple a grid-dip meter to a toroid inductor — and generally fail.

Until fairly recently ferrite and powdered-iron toroids were not available for frequencies above a few hundred kilohertz. Now, however, ferrites are available

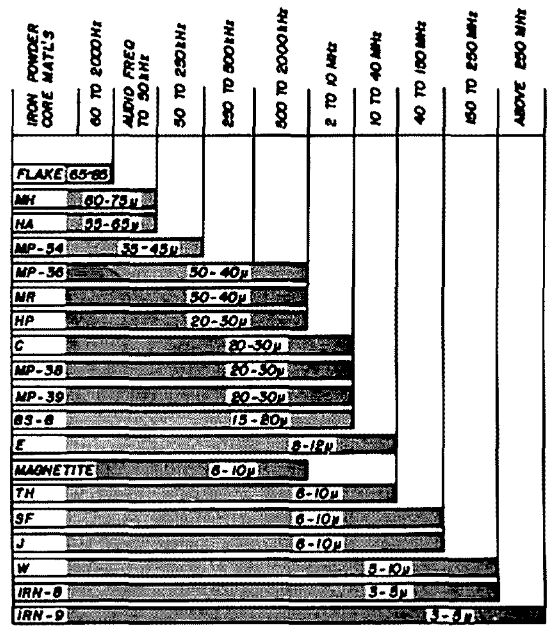


fig. 9. Frequency ratings for Arnold Engineering toroid-core materials.

that are useful into the GHz range, and powdered iron available for use up to several hundred MHz.

Let's look at a few examples. Amidon Associates* markets most of the powdered-iron toroids manufactured by Micro-Metals. They have toroids that are molded of types 2, 6, and 10 mixes, as designated by Micro-Metals. These mixes are all carbonyl types. Carbonyl is the chemical name for a compound formed of some substance and a carbon mon-

table 2. Identification of Micro-Metals core materials.

mix number	basic iron powder	color code
1	Carbonyl C	blue
2	Carbonyl E	red
3	Carbonyl HP	gray
4	Carbonyl J	blue and white
5	Carbonyl L	brown
6	Carbonyl SF	yellow
7	Carbonyl TH	white
8	Carbonyl GQ4	orange
10	Carbonyl W	black
12	IRN 8	green and white
13	IRN 9	green and red
15	Carbonyl GS6	red and white
19	electrolytic	—
22	flake	green and orange
41	hydrogen reduced	green
42	hydrogen reduced	—

*Amidon Associates, 12033 Otsego Street, North Hollywood, California 91607.

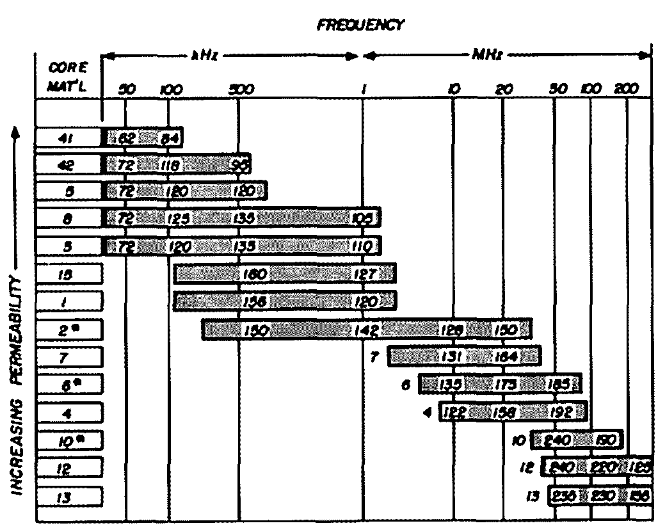


fig. 8. Frequency ratings for Micro-Metals toroid cores. Asterisked core-material mix numbers are available from Amidon Associates. Numbers on bars are Q readings.

table 3. Typical toroid-core winding data for Micro-Metals cores available from Amidon Associates.

core	wire size (AWG)	number of turns	inductance (μH)	frequency (MHz)	Q
T-94-2	30	125	130.0	0.90	232
	15/44 litz	200	328.0	0.78	278
	15/44 litz	400	1420.0	0.37	276
T-80-2	20	36	7.8	4.0	280
	26	80	37.0	2.5	246
	34	220	276.0	0.8	188
T-68-2	20	26	3.9	5.5	260
	28	79	33.0	2.5	240
	34	197	192.0	1.0	190
T-50-2	20	19	2.08	6.4	207
	30	79	33.0	2.3	200
	32	200	218	0.4	124
T-37-2	20	12	0.64	8.0	158
	24	22	2.16	7.0	170
	26	28	3.34	6.0	183
T-25-2	26	14	0.72	12.0	136
	30	30	3.22	8.0	162
	36	65	14.5	5.0	148
T-12-2	28	9	0.19	21.0	75
	32	17	0.65	15.0	84
	40	40	3.37	10.0	85
T-94-6	16	25	4.7	5.0	350
	20	20	3.0	6.0	340
	20	35	8.7	3.0	339
T-80-6	16	20	1.88	9.0	317
	20	15	1.1	10.0	255
	20	28	3.6	6.0	299
T-68-6	20	23	2.42	10.0	304
	20	15	1.08	10.0	270
	22	33	5.1	6.0	305
T-50-6	18	14	0.86	14.0	252
	22	25	2.60	10.0	260
	26	42	7.5	6.0	244
T-37-6	20	12	0.48	18.0	181
	22	17	0.97	14.0	194
	26	28	2.45	10.0	195
T-5-6	24	10	0.30	21.0	152
	28	20	1.10	15.0	164
	36	67	11.7	6.0	138
T-12-6	30	11	0.23	25.0	92
	34	18	0.56	20.0	90
	36	25	1.06	15.0	96
T-50-10	20	10	0.37	25.0	178
	20	15	0.81	18.0	190
	22	20	1.38	13.0	188
T-37-10	20	8	0.20	30.0	138
	22	15	0.61	20.0	165
	26	25	1.54	15.0	162
T-25-10	22	7	0.12	45.0	135
	24	9	0.18	35.0	141
T-12-10	28	7	0.06	60.0	120
	30	11	0.16	40.0	101
	32	14	0.26	35.0	87

oxide molecule; all three ferromagnetic metals form such carbonyls: $Fe(CO)_5$, $Ni(CO)_4$ and $Co_2(CO)_8$.

The Micro-Metals cores available from Amidon Associates, mixes 2, 6, and 10, are identified as carbonyl E, carbonyl SF, and carbonyl W. The same carbonyls are listed by Arnold Engineering Company in their catalogs. **Fig. 8** shows the general frequency range of usefulness for Micro-Metals materials, and a similar chart published by Arnold is shown in **fig. 9. Table 2** identifies the Micro-Metals mix number with the more commonly used name; e. g., carbonyl E and mix number 2 are the same. As you can see the graphical representations are in general agreement; differences reflect details of material use.

Table 3 shows the performance of toroid cores available from Amidon Associates, and should give you a good feel for which core to use at a given frequency. These point values were apparently extracted from Micro-Metals' "Engineers Aide Handbook" which is more complete, giving continuous curves of Q vs frequency for each core with different numbers of turns. This handbook is very nice, but it costs $10.00; **table 3** will probably suffice for most amateur designers. **Fig. 10** is an example of how the Micro-Metals handbook treats a typical core.

The surprising thing about coils wound on powdered-iron toroids is their Q. Note that Qs of several hundred are quite typical. The only other coils that give such high Q are air-wound types such as B&W Miniductor or Airdux (and then only when they're not too close to the chassis or in special high-conductivity shield cans).

Toroids are also available in ferrite material. You will find that ferrite toroids provide coils that have much larger inductance for a given number of turns because the ferrites have a much higher effective μ then the powdered irons. For comparison purposes consider a Micro-Metals T-44-6 which has an inductance of 0.5 μH for 10 turns. A similar sized ferrite core of Q1 material (Indiana General CF103) with the same number of turns will provide

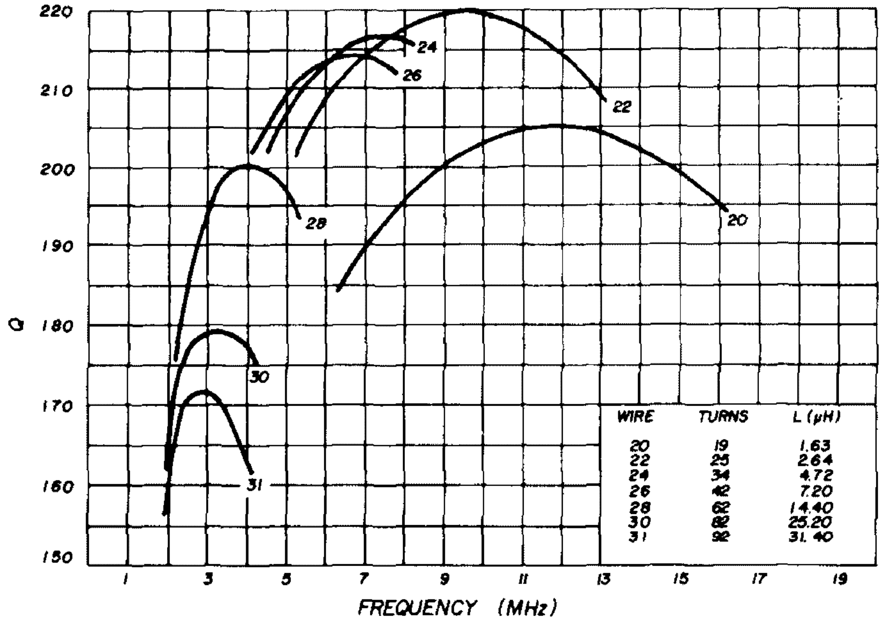


fig. 10. Winding data for Micrometals T-50-6 toroid core.

table 4. Performance of Amidon balun kit as measured with Boonton 250A RX meter. For this data a 200-ohm carbon resistor was used as a balanced load.

frequency (MHz)	resistance (ohms)	parallel X_L
3.5	53.8	+j90
7	53.0	+j112
14	52.5	+j204
29	49.6	+j220
50	44.5	+j122

inductance of 3 μH.

This fact would seem to be all on the positive side; large inductance coils with few turns. But it's not all "beer and skittles!" Since one turn difference in the coil can cause such a large change in inductance, it is difficult to achieve an *exact* inductance value (there's no such thing as a half turn in a toroid coil). Also, it turns out that ferrite coils have a rather severe variation of both inductance *and* Q with temperature.

The ferrite cores that are most easily obtainable in small quantity are Indiana General types. A good representation of the Indiana General catalog is carried by Newark Electronics in their 1971 mail-order catalog.

One of the more popular applications for large diameter toroid cores is as balanced-to-unbalanced transformers (baluns) for antennas. The Amidon balun kit is a typical example. The performance of this kit, connected as a 4:1 balun, is shown in table 4 (balun winding consisted of 14 turns bifilar-wound number-14 wire).

ferrite-core rf choke

Ferrite rods are commonly used as cores for high-current rf filament chokes. A typical example is the filament-choke kit available from Amidon Associates that uses a 12 X 100 mm ferrite-rod core. With the two bifilar windings connected in parallel (as used in a typical circuit) it has 26-μH inductance. This represents 500-ohms at 3.5 MHz, or ten times the usual 50-ohm driving impedance.

To test the choke I mocked up a typical linear amplifier using two grounded-grid 811As in parallel. Total capacitance to grid was 23 pF; this gives a parallel resonance at 6.5 MHz that was confirmed with a grid-dip meter. Between

table 5. Characteristics of the Amidon Associates ferrite-core filament choke.

frequency (MHz)	parallel capacitance (pF)	parallel resistance (ohms)	equivalent parallel reactance
7	17.5	∞	−j1500
14	18.7	35k	−j600
21	23.6	25k	−j330
28	24.5	25k	−j240

table 6. International standardized pot-core sizes.

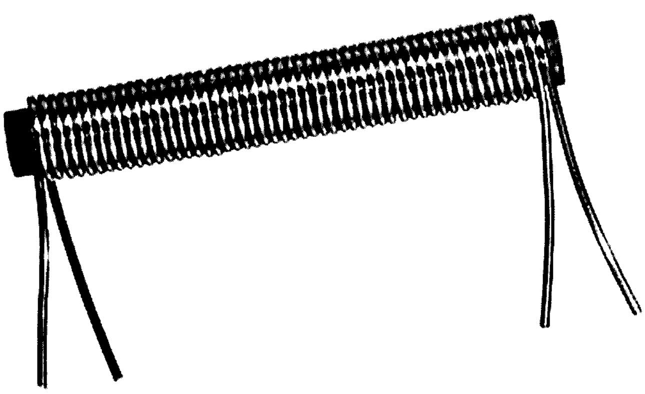
diameter (mm)	height (mm)	Siemens part	Indiana General part	TDK part	Nippon Electric part	Magnetics Inc. part	Ferroxcube part
9	5	65 521	--	--	--	40905	--
11	7	65 531	--	--	--	41107	1107P
14	8	65 541	TC1-01	P14/8	P14/8	41408	1408P
18	11	65 551	TC1-02	P18/11	P18/11	41811	1811P
22	13	65 661	TC1-03	P22/13	P22/13	42213	2213P
26	16	65 671	TC1-04	P26/16	--	42616	2616P
30	19	65 701	TC1-05	P30/19	--	43019	3019P
36	22	65 611	TC1-06	--	--	43622	3622P

6.5 and 30 MHz I could find no parallel resonant points with the grid dipper. Performance of the filament choke, as measured with a Boonton RX meter, is shown in table 5.

pot cores

To cut down the high effective μ and temperature coefficient of ferrite-wound coils, they are usually wound on gapped cores. The most common gapped ferrite core is the pot core. This form is particularly nice because the gap is in the center post where it does not couple to the outside world. Furthermore, most pot cores have a slug available that can be adjusted in the gap to vary inductance by ± 5%. Another advantage is gained with pot cores: The core comes in two matched halves; the coil form may be easily wound and then inserted into the core halves. There are various clamping arrangements for holding pot-core halves together; each manufacturer has his own mechanical technique.

Amidon Associates ferrite-core rf choke.



Fortunately, there has been some standardization in pot-core sizes. Since they seem to be of European origin the sizes are all in millimeters. Table 6 shows the standard pot core sizes. Also, a uniform method of designating what inductance a gapped pot core of so many turns will produce has been adopted by manufacturers. The AL core number designates nanohenries per turn-squared. A nanohenry is 1/1000 of one μH; the rating is per turn *squared* because induc-

using the pot-core winding tables

Assume the required inductance, L_X, is 150 mH. A ferroxcube pot core, 1811PA100 is to be used. Find the number of turns required, and the largest usable wire size.

1. Number of turns $= \frac{L_X}{A_L} \times 1000 = \frac{150}{100} \times 1000 = 1225$ turns
2. Winding area of bobbin of 0.030 square inch (see table 5).
3. Turns per square inch: $\frac{1225}{.030} = 40{,}800$
4. From table 6, 39 AWG, heavy-Formvar insulation, has 53,855 turns per square inch, thus can be used to wind the 150-mH inductor.

Assume a 22-mm core, Ferroxcube 2213P-A250 is wound with 40 AWG single Formvar. Calculate inductance when bobbin is fully wound.

1. Winding area for bobbin is 0.047 square inch (see table 5).
2. Turns per square inch for 40 AWG single Formvar is 82,180 (see table 6).
3. Winding factor of 90%: (82,190 X 90%) = 74,000 turns per square inch.
4. Number of turns: (74,000 X 0.047) = 3478.
5. Inductance is $\mu_e L_o$ mH (see table 5 for values). Therefore: (97.6) X (2.50 X 10^{-6}) X (3478 turns) = 2940 mH.

table 7. Abbreviated pot-core data for Ferroxcube cores. This information is used with the wire-winding data in table 6.

core diameter	pot core part no.	available core material 3B5	3D3	3H1	induct. (m h)/ 1000N)	μ_e	winding area (sq. in.) single section bobbin	LO (mHO)	accessories Bobbins	Hardware
14 mm	1408P-A40*	-	X	-	40	24	0.0158	$1.66N^2 \times 10^{-6}$	1408 F1D	1408H
	1408P-A60*	X	X	X	60	36			1408 F2D	
	1408P-A100*	X	X	X	100	60				
	1408P-A160*	X	-	X	160	96				
18 mm	1811P-A40*	-	X	-	40	18	0.030	$2.19N^2 \times 10^{-6}$	1811 F1D	1811H
	1811P-A60*	X	X	X	60	27			1811 F2D	
	1811P-A100*	X	X	X	100	45			1811 F3D	
	1811P-A160*	X	X	X	160	73				
	1811P-A250*	X	-	X	250	114				
22 mm	2213P-A60*	-	X	-	60	23.4	0.047	$2.50N^2 \times 10^{-6}$	2213 F1D	2213H
	2213P-A100*	X	X	X	100	39			2213 F2D	
	2213P-A160*	X	X	X	160	62.5			2213 F3D	
	2213P-A250*	X	-	X	250	97.6				
	2213P-A400*	X	-	X	400	156				
	2213P-A600*	X	-	X	600	234				
26 mm	2616P-A100*	-	X	-	100	31	0.067	$3.25N^2 \times 10^{-6}$	2616 F1D	2616H
	2616P-A160*	X	X	X	160	49			2616 F2D	
	2616P-A250*	X	-	X	250	77			2616 F3D	
	2616P-A400*	X	-	X	400	123				
	2616P-A600*	X	-	X	600	185				

*Add material designation to complete part number

tance increases as the square of coil turns. The AL number is almost always printed on the flat end of a pot core and is easy to find. If the core has no air gap it is usually marked OL.

As an example of using the AL number assume a Siemens core of N22 material; 18 X 11 mm with an AL number of 160. With 100 turns of wire on it the expected inductance will be:

$$L = (AL)\,T^2 = (160\text{ nH})\,(100T)^2 = 1.6 \times 10^6\text{ nH} = 1600\,\mu\text{H or } 1.6\text{ mH}$$

The size of the wire should be chosen to fill the bobbin, and charts are available to help choose proper wire size. Typical charts are shown in tables 7 and 8. These charts are for Ferroxcube pot cores, but since pot cores come in standard sizes they can be used approximately for other brands. (Siemens has a chart for metric wire sizes which would be of interest to European amateurs.)

There are a number of different ferrites from which pot cores can be used. Each manufacturer has his own designation for these and a recommended frequency range for each. Ferrites are made in this country by Indiana General, Stackpole, Allen Bradley, Ferronics and Arnold Engineering. Foreign manufacturers that market in the U. S. are Siemens, Ferroxcube (Phillips), Nippon Electric and TDK. Since each company has its own set of materials and designations, it is a bit hard to present a list of what type is best for a given frequency.

broadband rf transformers

Until ferrite and high-frequency powdered-iron cores became available there was no such thing as a broadband rf transformer. Now, thanks to these materials, they are widely used from below hf to above vhf. These broadband transformers use closed magnetic structures such as toroids, non-gapped pot cores and

table 8. Wire-winding data shows maximum number of turns of various wire sizes for the pot cores shown in table 5. Since pot cores come in standard sizes this data may be applied approximately to all brands. This information is based on exact layer winding. Normally 90% to 95% of these values are used, depending on whether the coil is random or layer wound.

wire size AWG	diameter	cross-sectional area circular mils	cross-sectional area square inches	turns per sq inch hf wire	turns per sq inch sf wire	resistance per 1000 ft
15	.0571	3260	256 10^{-5}	259	275	3.18
16	.0508	2580	203 10^{-5}	327	346	4.02
17	.0453	2050	161 10^{-5}	407	432	5.05
18	.0403	1620	128 10^{-5}	509	544	6.39
19	.0359	1290	101 10^{-5}	634	679	8.05
20	.0320	1020	804 10^{-6}	794	854	10.7
21	.0285	812	638 10^{-6}	989	1063	12.8
22	.0253	640	503 10^{-6}	1238	1343	16.2
23	.0226	511	401 10^{-6}	1532	1677	20.3
24	.0201	404	317 10^{-6}	1893	2094	25.7
25	.0179	320	252 10^{-6}	2351	2632	32.4
26	.0159	253	199 10^{-6}	2932	3326	41.0
27	.0142	202	158 10^{-6}	3711	4112	51.4
28	.0126	159	125 10^{-6}	4581	5213	65.3
29	.0113	128	100 10^{-6}	5621	6383	81.2
30	.0100	100	785 10^{-7}	7060	8145	104
31	.0089	79.2	622 10^{-7}	8455	10,097	131
32	.0080	64.0	503 10^{-7}	10,526	12,270	162
33	.0071	50.4	396 20^{-7}	13,148	15,615	206
34	.0063	39.7	312 10^{-7}	16,889	19,654	261
35	.0056	31.4	246 10^{-7}	21,163	25,531	331
36	.0050	25.0	196 10^{-7}	26,389	31,405	415
37	.0045	20.2	159 10^{-7}	31,405	39,570	512
38	.0040	16.0	126 10^{-7}	39,567	49,070	648
39	.0035	12.2	962 10^{-8}	53,855	65,790	847
40	.0031	9.61	755 10^{-8}	65,790	82,180	1080
41	.0028	7.84	616 10^{-8}	-	98,856	1320
42	.0025	6.25	491 10^{-8}	-	121,174	1660
43	.0022	4.84	380 10^{-8}	-	158,246	2140
44	.0020	4.00	314 10^{-8}	-	205,517	2590
45	.0018	3.24	254 10^{-8}	-	249,855	3200
46	.0016	2.56	201 10^{-8}	-	310,205	4050

note: hf is heavy Formvar insulation; sf if single Formvar.

multi-aperture cores. The amateur radio literature has a number of references on using toroids as balance-to-unbalance transformers (to match 50-ohm coax to balanced antennas).[7,8,9] Most of these references are derived from an article in the *Proceedings of the IRE*.[10] Balun-type transformers have become so popular that Amidon Associates offers the necessary toroid cores and wire in a balun-kit. The references show how to build 1:1 and 4:1 transformers (either balanced-to-unbalanced or unbalanced-to-unbalanced).

By using high-frequency ferrite pot cores, North Hills has rf transformers that will match 50- or 75-ohm coax to balanced impedances from 50 to 800 ohms. These transformers are apparently more difficult to build than the 1:1 or 4:1 types; they are probably adjusted at the factory with time-domain reflectometry (TDR). Reference 10 gives some details on optimizing bandwidth using TDR techniques. In using the North Hills transformers I found that a 50-to-200-ohm unit will not function well as a 75-to-300-ohm transformer, and so on. This is because the bifilar winding apparently acts as a transmission line which shouldn't be mismatched.

Broadband rf transformers are now available from at least four firms, in-

cluding North Hills, Vari-L, Relcom and Vanguard in a variety of ratios. North Hills even has one that will take 1000 watts.

The broadband rf transformer has made a number of new devices possible. Probably the most important of these is the double-balanced hot-carrier diode modulator. These wonderful devices are available commercially from a number of

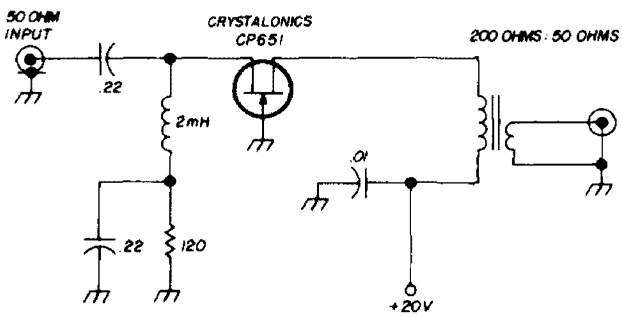


fig. 11. Basic circuit of broadband fet rf preamplifier used in Comdel HDR101. Bandwidth of unit is 0.5 to 40 MHz, voltage gain is 9 dB, noise figure is 2.5 dB, dynamic range is 140 dB.

firms. They are the nearly perfect mixer that was wanting for so many years. References 11, 12 and 13 cover some of the details and use of these mixers, and reference 4 shows how they can be built at home.

Comdel, Inc.,* is using broadband rf transformers and fets to build a very fine high-frequency preamplifier. The Comdel HDR-101 is nearly perfect for improving sensitivity of communications receivers with minimum decrease in maximum signal handling capacity.[14] A typical high-frequency communications receiver may have a noise figure of 12 dB; the Comdel preamp has a noise figure of 2.5 dB with 9-dB gain. The combination yields a receiver system having 9 dB more gain and a 5.5-dB noise figure. A general description of the Comdel circuit was published in references 15 and 16; it is shown in **fig. 11.** Since the output transformer can be handwound as described in reference 10 this circuit is within the realm of home construction. Care is necessary in choosing the input rfc. No doubt there are a number of subtle engineering tricks that contribute to the excellent performance of the Comdel unit, but the circuit of **fig. 11** can serve as a start for the serious experimenter.

An interesting circuit for a broadband push-push doubler using a North Hills 50-to-400-ohm unbalanced-to-balanced transformer* is shown in **fig. 12.** The circuit cancels the fundamental by *balance* and requires no filtering to produce more second harmonic than fundamental. While it is shown using a matched pair of fets, I obtained somewhat similar results with a selected pair of HEP802s chosen with nearly equal I_{dss}.

The 400-ohm balanced to 50-ohm ferrite unbalanced transformer used in **fig. 12** is also quite nice for adapting older hf receivers to newer 50-ohm systems. Most older hf receivers (such as my Hallicrafters S20R) have a 400-ohm balanced antenna input which was designed to match the Zepp and similar

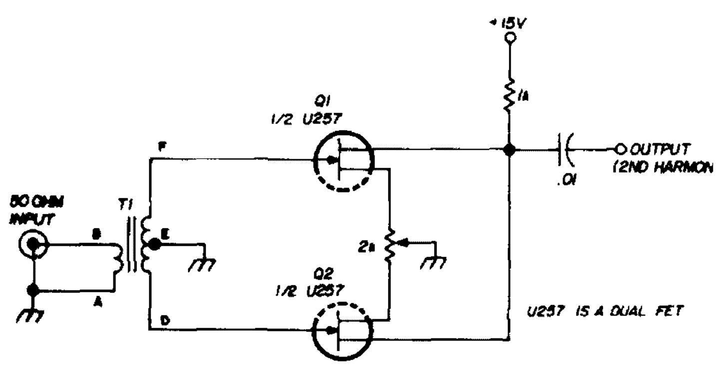


fig. 12. Broadband push-push doubler circuit. T1 is broadband 50-ohm-unbalanced to 400-ohm balanced transformer (North Hills 556026-336).*

antennas of the era. If you're one of the fortunate few who have a large rhombic fed with 600- or 800-ohm open-wire line commercial broadband transformers are available that will readily adapt your antenna system to a 50-ohm unbalanced receiver.

*The author has a limited number of these transformers available to experimenters at $5.00 plus $0.50 for postage and packing. Write to Hank Olson, Post Office Box 339, Menlo Park, California 94025.

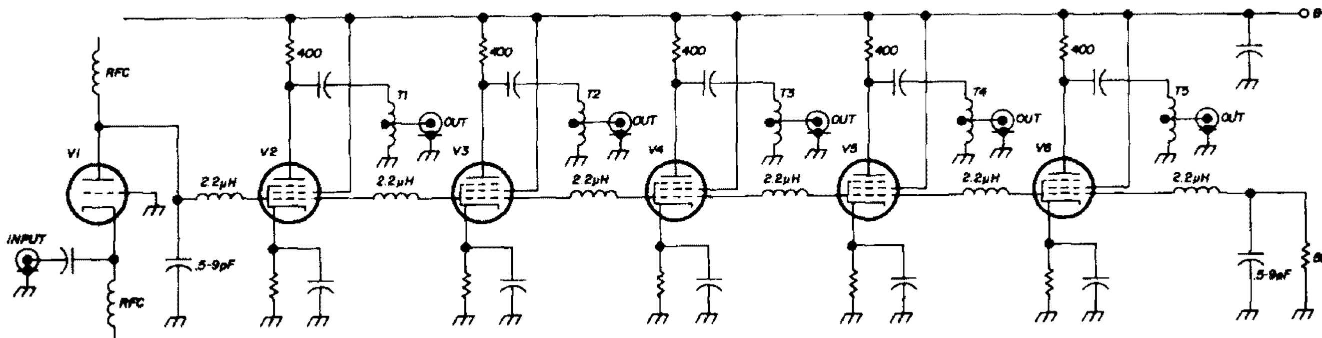


fig. 13. High-frequency multicoupler isolates receivers from antenna. Transformers are 200-ohm-to-50-ohm balanced-to-unbalanced, such as the North Hills 0502AA. Tube V1 is a high-transconductance triode (20,000 umho); other tubes are sharp-cutoff pentodes.

By combining a distributed amplifier with a broadband ferrite transformer it is a fairly simple matter to make a multicoupler as shown in **fig. 13**; it is used to connect up to five hf receivers to the same antenna. Each receiver is isolated from the others and may be used at any frequency in the hf band (assuming that the antenna is a broadband type such as a log-periodic). Note that the input capacitance of the pentodes and the 2.2-μH inductors from an artificial lumped-constant 500-ohm transmission line – this is why the input is broadband. This vacuum-tube design is obsolete in the semiconductor age but modification to fets should be fairly straightforward.

There are numerous uses for ferrite and powdered-iron cores in measurement equipment. Doug DeMaw's low-power directional power meter is a good example.[17] Another good antenna-measuring device uses a two-hole ferrite shielding bead in a bridge.[18] This simple bridge can be built with a two-hole ferrite bead available from W6KZK.

summary

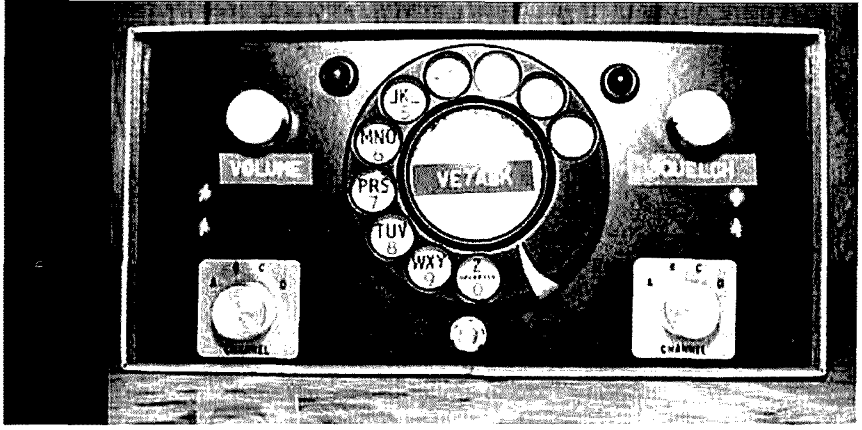
High-frequency ferrites and powdered iron enable us to use techniques at higher frequencies which formerly were restricted to audio and power frequencies. If you can think of a magnetic component that is used at audio, the chances are quite good that the same technique can be applied at hf or vhf with the cores that are currently on the market.

*The ferrite core with windings, as described in reference 18, is available for $.50 from Swan Antenna Company, Post Office Box 1122, Stockton, California 95201.

bibliography

1. O. Heaviside, "Electromagnetic Theory," Volume 1, 1893, p. 441.
2. H. Olson, "Remember the Blocking Oscillator," *Electronic Products,* December, 1968, p. 54.
3. H. Olson, "Blocking Oscillators," *ham radio,* April, 1969, p. 45.
4. W. Ress, "Broadband Double-Balanced Modulator," *ham radio,* March, 1970, p. 8.
5. T. Marshall, "Development of Integrated Ferrite Bead RF Attenuators and Dissipative RF Filter Providing High Attenuation at Low Frequencies," Technical Memorandum 1630, Picatiny Arsenal, June, 1965 (DDC number 463908).
6. E. Wetherhold, "Inductance and Q of Modified Surplus Toroidal Inductors," *QST,* September, 1968, p. 36.
7. R. Turrin, "Broadband Balun Transformers," *QST,* August, 1959, p. 33.
8. D. DeMaw, "Toroid-Wound Inductors," *QST,* January, 1968, p. 11.
9. W. Orr, "Broadband Antenna Baluns," *ham radio,* June, 1968, p. 6.
10. C. Ruthroff, "Some Broadband Transformers," *Proceedings of the IRE,* August, 1959, p. 1337.
11. Hewlett Packard, "Double Balanced Mixers," Technical Data Sheet, 15 January, 1968.
12. Relcom, "Double Balanced Mixer," Technical Data Sheet, Model M1, January, 1970.
13. J. Fisk, "Double-Balanced Modulators," *ham radio,* March, 1968, p. 48.
14. New Apparatus, "Comdel HDR-101ac Broadband RF Amplifier," *QST,* August, 1969, p. 23.
15. "Field-Effect Transistor Key to 140-dB Dynamic Range in Broadband Amplifiers," *Communication Designer's Digest,* April, 1968, p. 10.
16. Crystalonics, Technical Memo TMF-1 (part of Annual Catalog, Fall, 1968, p. 102).
17. D. DeMaw, "The QRP 80-40 C. W. Transmitter," *QST,* June, 1969, p.16.
18. O. Swan, "Impedance Bridge," *ham radio,* February, 1970, p. 66.

ham radio



customizing the fm control head

Modifications to improve performance of older mobile fm sets

Vern Epp, VE7ABK, 203 View Street, Nelson, B. C., Canada

Many two-meter fm enthusiasts can't afford the latest Motrac or Motran equipment for mobile operation and must be content with older offerings. Many older vibrator-powered mobile sets are available at reasonable prices but have certain disadvantages.

Some older sets have a frequency selection of four channels; however most are limited to one or two channels. Also, most older sets are trunk-mounted, which requires a control head. These heads aren't too attractive, especially when used in late-model cars. Furthermore, when it's desired to add other circuits, these heads can't be used unless they're large.

In this article I've described some modifications that will improve a typical two-meter mobile fm set – the RCA Carfone. These modifications are adaptable to other sets with a little ingenuity.

The original Carfone remote head was reinstalled in a new enclosure for enhanced appearance. Other features were also added to make this equipment competitive with late-model mobiles:

1. Modification from 2- to 5-channel operation.

2. Modification for a dynamic microphone.

3. Addition of a dial encoder for repeater and auto patch.

4. Adaptability for channel scan (which will be added later).

multichannel modifications

A mobile fm set can be modified in many ways for multichannel operation. If space is a consideration, it is advantageous to remotely locate the transmit and receive crystal oscillators. The circuit of fig. 1 was added to the control head of my set to increase frequency coverage. (Interested readers may wish to read the article in which this circuit first appeared.)[1]

Referring to fig. 1, a multichannel crystal oscillator and emitter follower provide signal inputs to the trunk-mounted set via RG-58/U coax cable. Selector switches for transmit and receive modes are independent, so any combination of frequencies within the limits of the circuit is possible. This is especially desirable because many frequencies are used on two meters for repeaters with simplex operation. Printed circuit boards are available for those wishing to make this conversion.*

Two circuit boards are required, one for the transmit and one for the receive channel. The transmit-channel oscillator supplied sufficient drive to produce the

*Write Sam Craig, W2ACM, 5812 Tilton Road, East Syracuse, New York 13057.

same output as that from the original oscillator in my unit. However, the injection signal from the receiver board was rather low, so I included a resonant circuit consisting of a slug-tuned inductor leaving one set of crystals in the trunk-mounted unit.

microphone modification

Most older fm sets use a carbon mike.

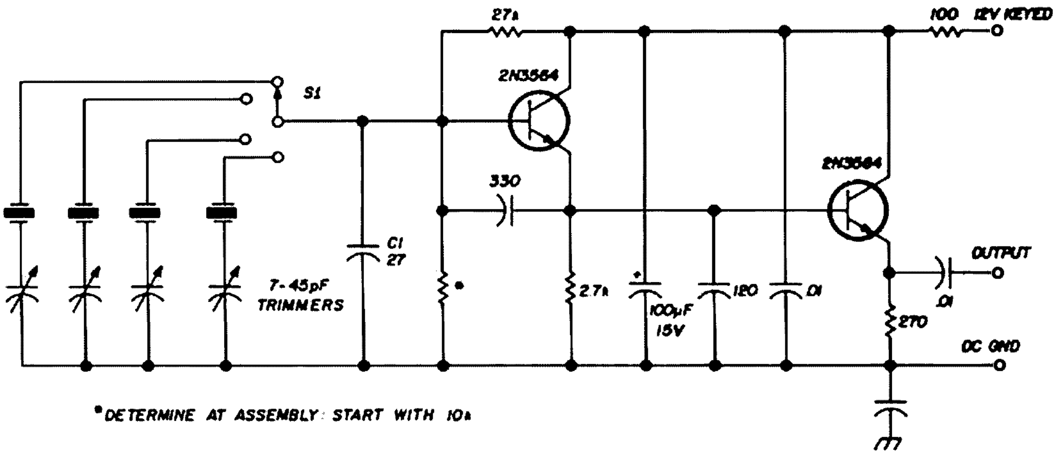
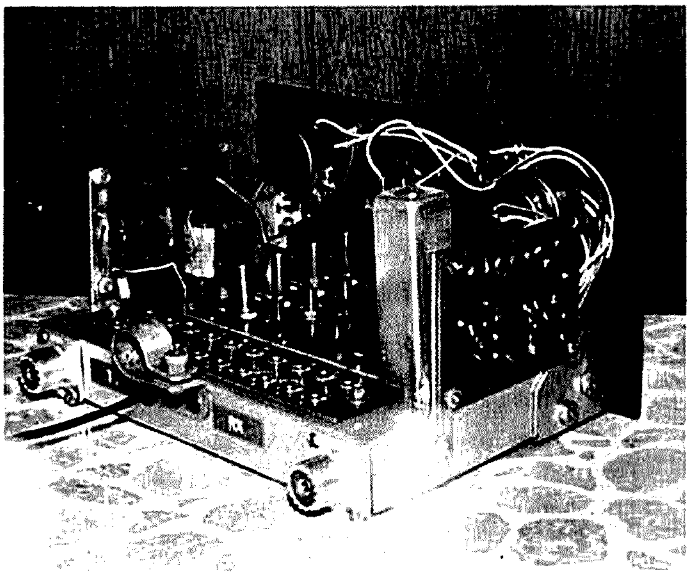


fig. 1. Remote oscillator circuit for multifrequency operation. Two circuit boards are required. Receive board is fed directly from 12 Vdc at on-off switch. Transmit board is fed by a keyed 12 Vdc source. Output of each is fed via RG-58/U coax to trunk-mounted set between grid and ground of original oscillators.

and parallel capacitor tuned to the crystal frequency. This circuit was installed between the receiver-board output and the 0.01-μF coupling capacitor (fig. 1). The signal was then routed via RG-58/U cable to the set in the car trunk.

If your mobile set has two channels, with this addition you'll have five channels. The remote unit supplies four channels, and a fifth is obtained by

Rear view of the RCA Carfone showing addition of transmit and receive oscillator boards for extended frequency coverage.



Its audio quality is fair but not the best. The modern (medium- to high-priced) fm mobile units are designed to use a dynamic mike that has a built-in preamplifier.

My unit uses a Motorola CTMN6019A mike, which requires approximately 4 volts negative to drive the built-in preamp (fig. 2). Some mikes, like the Shure 405T, operate from a positive-supply voltage, which is readily available from the B+ supply through a voltage divider and filter. The improvement in audio quality is remarkable, and the new microphone is much more attractive next to the new control head.

dial and encoder

The telephone dial, which is mounted on the front panel of the modified control head, pulses the tone encoder. My encoder unit, shown in fig. 3, oscillates at about 2.6 kHz. The oscillation frequency is easily changed by selection of the capacitor across the toroid transformer primary. Any good npn transistor can be used. Component values aren't at all critical. The encoder frequency will have to fit the requirements of your area.

This circuit can be used for auto patch

control with a repeater station that includes this feature. Auto patch is becoming quite popular in the U. S. as well

head as described is not equipped for this feature but is compatible with such a modification.

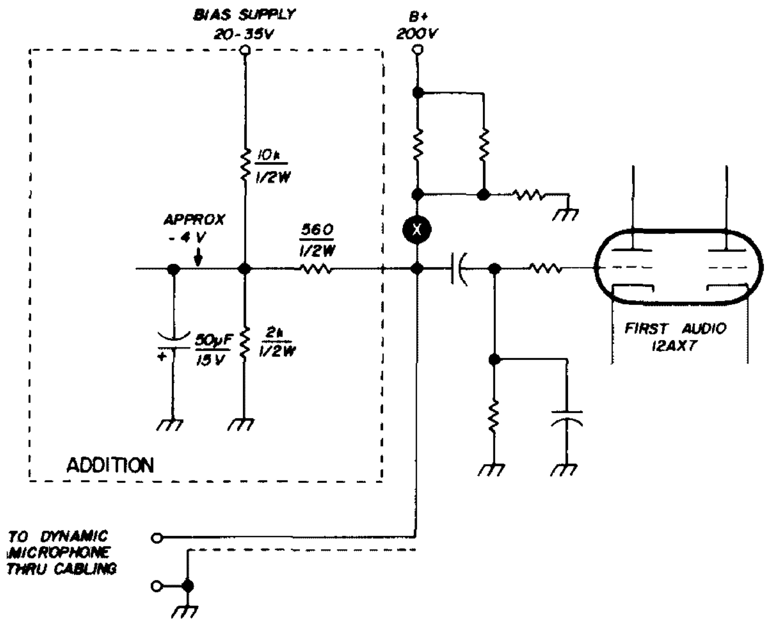


fig. 2. Modifications for converting the RCA Carfone mobile to a dynamic microphone input. Break circuit at point X.

as in Canada. It allows the mobile fm operator to communicate via landline telephone circuits and is sanctioned by "Ma Bell."[2]

channel scan

This is a monitoring system that permits you to keep track of more than one channel at a time, while automatically locking onto a transmitting

closing note

Mobile fm operation is a very popular part of ham radio, and the number of converts to this communication mode is increasing at a tremendous rate. The benefits of mobile fm over CB channels are obvious. Almost every metropolitan area has one or more amateur fm repeater stations that receive the mobile signal and retransmit it at increased power, thus

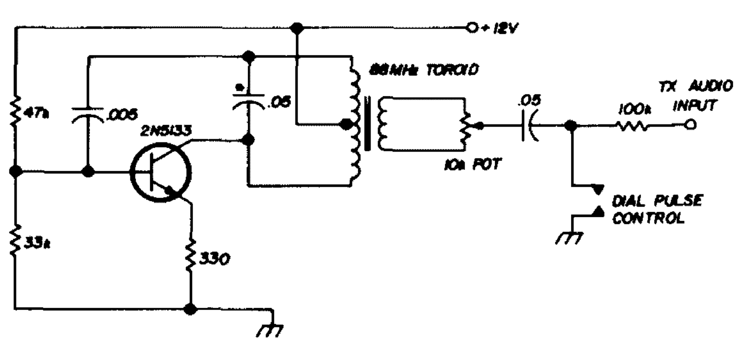


fig. 3. Dial encoder oscillator using an 88-mH toroid and junk-box parts. Capacitor marked with asterisk determines frequency.

channel. Most systems use a ring counter (one for each channel) triggered by a unijunction relaxation oscillator. The UJT, in turn, is operated by the noise rectifier of the mobile unit. The control

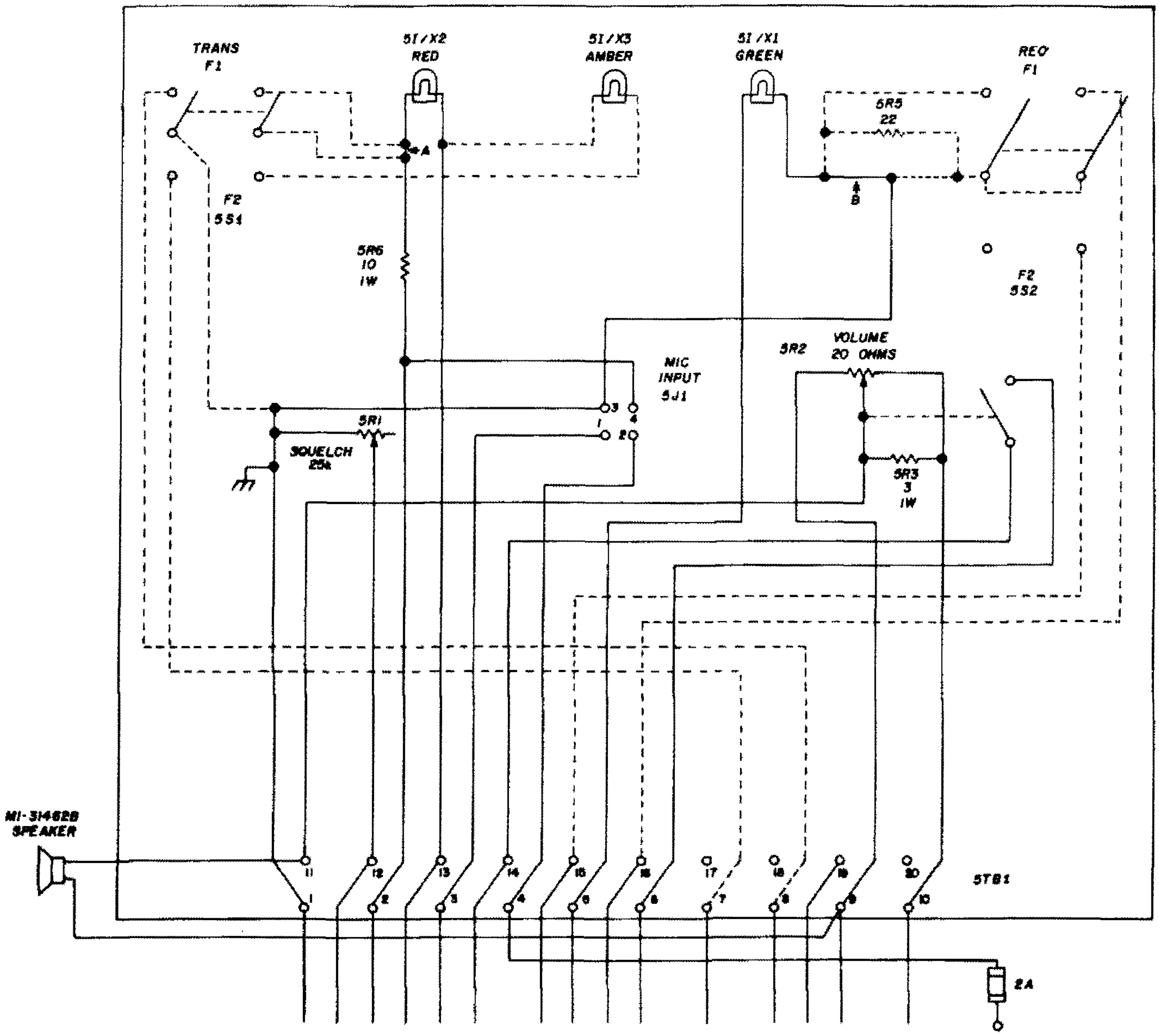
allowing even the most modest mobile signal to be "top dog" on the frequency.

Fm equipment is reliable and generally trouble-free. When it is necessary to service fm equipment, you won't have

much trouble getting your set back into operation because of the commonality of circuits — you can usually find someone who is familiar with your particular equipment.

your particular requirements. All circuits shown here have been tried and proven.

The improved appearance of the modified control head will keep the wife happy, and the dial will keep the kids



Circuit of original control head which was installed in the new control box.

Controls are simple and straight-forward, consisting of an on-off-volume control, squelch control, and push-to-talk microphone. Antennas are physically small and easy to install and tune.

I have described a number of ways to customize an older mobile two-meter fm installation. The extent of your modernization work will, of course, depend on

occupied on those long trips (but be sure the equipment is switched off!).

references

1. Sam Craig, W2ACM, "A Remote Multi-frequency Oscillator for Surplus FM Units," *73,* March, 1970, p.44.
2. Donald E. Chase, W0DKU, "The Wichita Autopatch," *73,* May, 1970, p. 118.

ham radio

power, voltage and impedance nomograph

These graphs permit rapid conversion from one unit of measurement to another — they are particularly useful when using an oscilloscope for waveform analysis

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When working with various electronic systems it's often necessary to think in terms of power, voltage, impedance and dBm, and to move rapidly from one unit of measurement to another. Rather than making a calculation each time you change from one unit to another, it is much easier to use the graphs in figs. **1** and **2**. These graphs are an analysis of power vs voltage as a function of impedance.

The chart in fig. 1 covers power levels from 1 milliwatt to 1 watt, while the graph in **fig 2** covers the range from 1 watt to 1 kilowatt. The power range in dB above a milliwatt, or dBm, is shown on the right hand side of each chart.

These graphs are extremely useful when using an oscilloscope for waveform analysis, and direct conversion to power level is necessary. They are also useful in the design and calibration of rf voltmeters and wattmeters.

how to use them

When laying out these graphs one of the prime considerations was to make them easy to use. All you have to do is enter the graph with the known quantity and continue to the appropriate impedance line; read the unknown value on the opposite axis of the graph. For example, what power level is represented by 10 volts peak-to-peak across a 50-ohm line? Enter the chart at the 10-volt point on the lower axis, project upward to the 50-ohm impedance scale, then to the left to 250 milliwatts. Note that this corresponds to 24 dBm on the right-hand side of the chart.

Although the impedance curves on the face of the chart are limited to the most common transmission-line impedances, the power and voltage curves are not restricted to these impedances. The impedance ruler at the top of each graph may be used to construct other impedance lines; simply draw a straight line

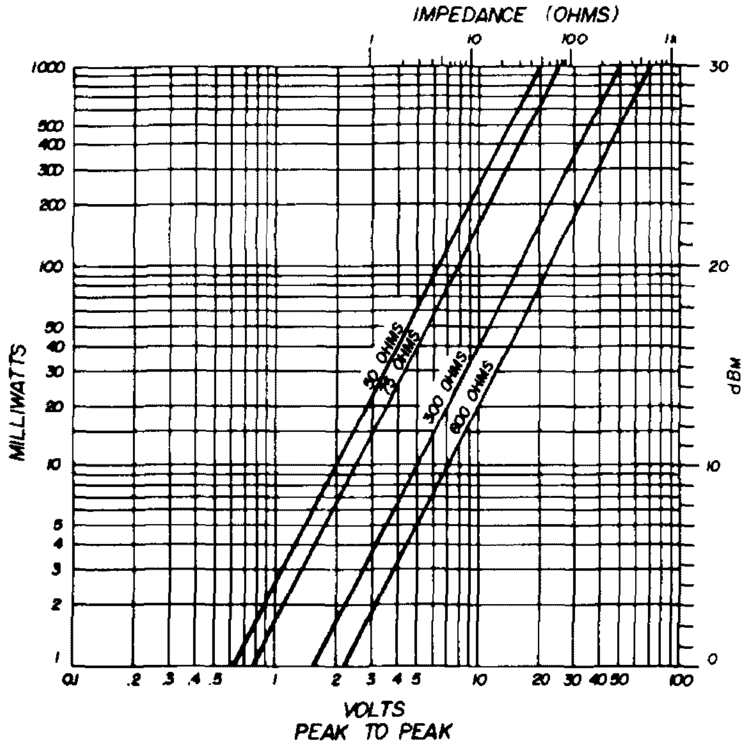


fig. 1. Power, voltage and impedance nomograph for power levels from 1 milliwatt to 1 watt.

through the appropriate point on the impedance ruler, keeping the new impedance line parallel to those already plotted.

The choice of peak-to-peak volts on the horizontal scale is based on the use of an oscilloscope as the primary measuring tool. If you wish, this axis can be recalibrated in any convenient terms that relate directly to peak-to-peak voltage. For example, for peak voltage, divide by 2; for rms voltage, divide the scale by 2.8.

I would appreciate hearing from any readers who find an unusual application for these graphs.

ham radio

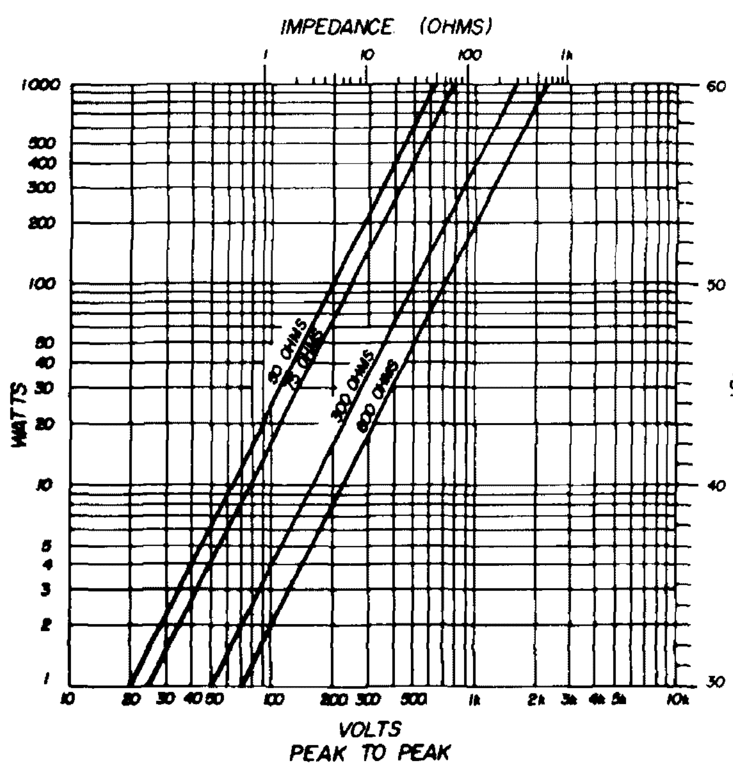
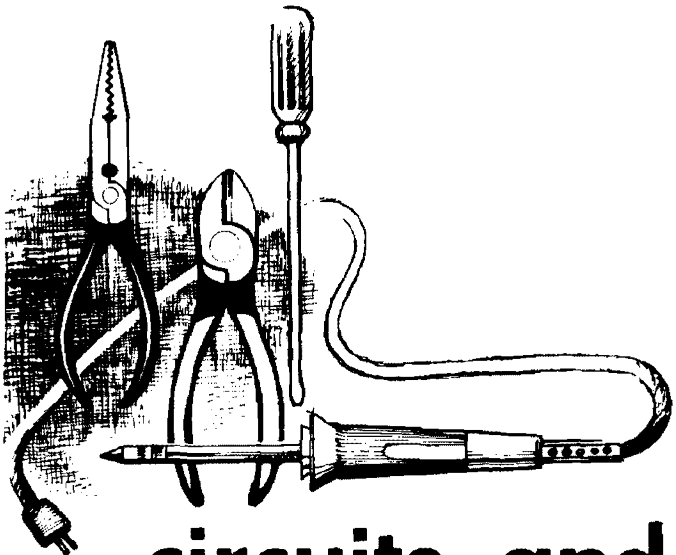


fig. 2. Power, voltage and impedance nomograph for power levels from 1 watt to 1 kilowatt.



circuits and techniques

ed noll, W3FQJ

power fets

Power field-effect transistors have now been developed and are available. Is it possible that power fets can be produced that will give bipolars and vacuum tubes a run for their money? Per watt cost remains high but is dropping. There is hope!

What are some of the advantages for power fets in high-frequency power-amplifier applications? Low input impedance and secondary breakdown problems, two headaches of bipolar transistors, are absent; interstage coupling and matching are easier. As an extra bonus there is low intermodulation distortion in linear rf amplification, less troublesome output circuit design and no thermal runaway. In addition, power fets are capable of high voltage operation so less current is required for a given dc input and rf output power.

fet characteristics

There are a number of important notations and parameters important to the operation of a field-effect transistor as a high-frequency power amplifier. These are listed in **table 1.**

In a field-effect transistor there is an essentially constant fixed relationship between pinch-off voltage, V_P, saturation current, I_{DSS}, and transconductance, g_{fs}. As gate bias is increased, drain current decreases in a square-law manner. In fact, the drain current changes between zero gate bias and pinch-off gate voltage in accordance with the following ratio:[1]

$$\frac{I_D}{I_{DSS}} = \left(\frac{V_P - V_{GS}}{V_P}\right)^2$$

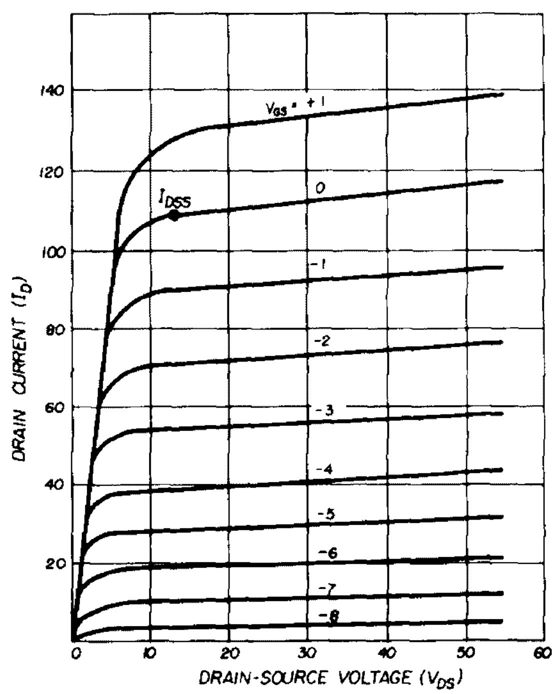


fig. 1. Typical field-effect transistor operating characteristics.

table 1. Field-effect transistor parameters.

I_D	dc component of drain current
I_d	instantaneous drain current
i_D	instantaneous total drain current (dc and ac drain current components)
I_{DSS}	drain current at zero gate voltage (note 1)
V_{DD}	drain supply voltage
V_{DS}	dc drain voltage
v_{ds}	instantaneous drain voltage
V_{GS}	gate-source voltage
V_P	gate-source pinch-off voltage (note 2)
v_{GS}	instantaneous total gate-source voltage (ac and dc components)
V_{PB} or	
V_{bias}	cut-off bias voltage
g_{fs}	common-source configuration forward transconductance
C_{iss}	common-source configuration input capacitance

note 1: I_{DSS} is usually measured at a low drain-source voltage, V_{DS}, corresponding to the beginning of the constant-current saturation region of the zero-bias curve, fig. 1.

note 2: The pinch-off voltage, V_P, is that gate-bias voltage, V_{GS}, at which the drain current is reduced to practically zero for a specified drain voltage, V_{DS}. This drain voltage is sometimes the one specified for the measurement of I_{DSS}.

From this expression you can develop the fundamental large-signal equations of a power fet.

class-AB linear operation

Above the I_{DSS} point on the zero bias curve (above pinchoff on all bias curves) the fet has a square-law transfer characteristic. If operation is confined to this range, intermodulation distortion components of any objectionable magnitude are not developed. Harmonic components do appear but these are suppressed in the output resonant circuit (fig. **2**).

Efficient operation and maximum linear output occur when the instantaneous peak drain current, i_D, reaches I_{DSS} at the positive crest of the input signal, and signal swing occurs over the entire range of the square-law region. Minimum load is placed on the signal source when the positive crest of maximum input signal swings no more positive than the zero gate voltage.

class-c operation

More output and higher efficiency can be obtained from the class-C mode. This can take advantage of greater drain voltage and drain current swings. Higher gate bias is used, and the conduction angle is approximately 120°.

However, somewhat more driving power is needed in class C. Input loading is light, especially when gate-channel conduction is avoided by not permitting the crest of the input rf wave to swing above zero gate bias.

In the practical planning of an fet rf power amplifier there are three equations of importance. These determine the required gate bias, zero-signal drain current and load impedance (see table **2**).

input circuit conditions

The load placed on the preceding stage by the input circuit is determined by the

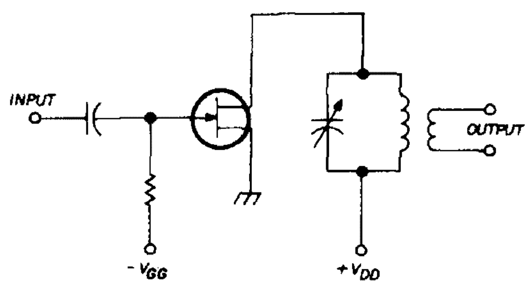


fig. 2. Basic fet tuned amplifier.

gate resistance, r_g and input capacitance, C_{iss}. Gate resistance is similar to the base resistance of a bipolar transistor and is determined by the resistivity of the semiconductor material; it acts in series with the common-source input capacitance. Since this capacitance is very low, input impedance is very high. Input impedance decreases with frequency because the reactance of C_{iss} decreases.

To summarize, most efficient operation without gate current occurs when

table 2. Design equations for fet rf power amplifiers.

Bias for class-AB linear	$V_{BIAS} = V_P/2$
Bias for class-C	$V_{BIAS} = V_P$
Load resistance Class-AB	$R_L = \frac{V_{DD} + (-V_P)}{I_{DSS} - I_D}$
Class-C	$R_L = \frac{V_{DD} + (-V_P)}{I_{DSS}}$
Dc drain current Class-AB	$I_D = \frac{I_{DSS}}{4}$
Class-C	$I_D = 0$

the peak of the rf drive signal reaches an instantaneous gate voltage that produces a drain current equal to I_{DSS}. The operation of a class-AB linear amplifier is maintained on the square-law portion of the transfer characteristic. To do so the operating bias must be set at a value of $V_P/2$. Based on the square-law response the static dc drain current for zero signal becomes $I_{DSS}/4$.

In class-C operation the conduction angle can be dropped to 120° and may swing off the square-law portion of the curve. In this case operating bias is set to the pinch-off value, V_P. Higher output is obtained and more driving power is needed. The zero-signal drain current becomes zero because of the pinch-off bias; drain current swings between zero and the I_{DSS} value.

Specifications for various high-frequency power field-effect transistors are given in table 3. Practical operating conditions for class-AB linear and class-C amplifiers can be calculated with the simple formulas shown in table 2.

Power fets are presently quite expensive, but useful experience can be obtained in the milliwatt level with less expensive types such as the Siliconix U-183 and Motorola HEP-801 and HEP-802. These devices cannot be classified as power fets but they permit operation up to 100 milliwatts.

A practical fixed-bias amplifier is shown in fig. 3. Note its similarity to vacuum-tube circuitry. Input and output resonant circuits are used. Gate neutralization is often needed because of the high input impedance of the stage. Inasmuch as the feedback capacitance Crss is reasonably constant, neutralization is usually possible with a *fixed* neutralizing capacitor. Typical gain on the 10-meter band is 20 dB in class AB and 25 dB in class C.

self-biasing

Power field-effect transistors can be operated as self-biased amplifiers. In this case the positive sweep of the drive signal is adjusted so that the gate-channel junction is forward biased at the peak of the gate input signal. Electron charges then flow instantaneously from the channel to the gate, and a negative dc bias can be developed with a gate-input resistor-capacitor network in fig. 4.

Input impedance is lowered with this system, and more driving power is required to overcome the gate circuit losses. However, you can dispense with an external gate-bias source. Although interstage coupling and matching are somewhat more troublesome than for the no-gate-current mode of operation, the problems are less difficult than those encountered in a bipolar amplifier of comparable power.

power oscillator

Power field-effect transistors perform

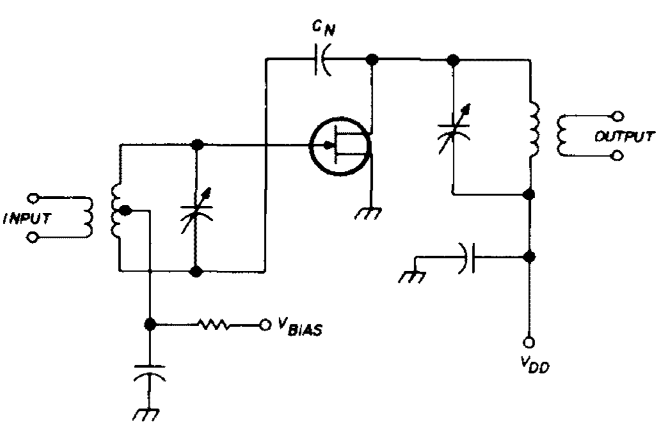


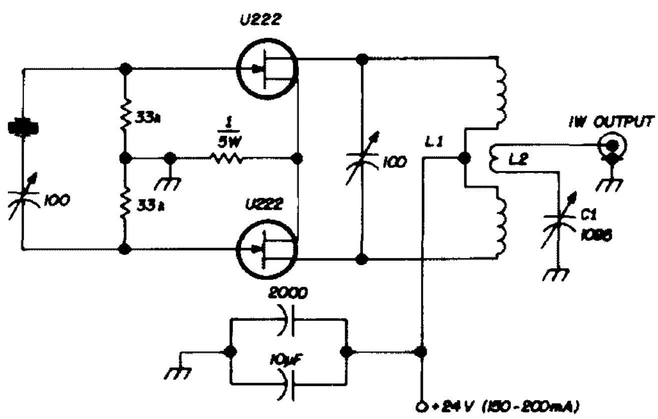
fig. 3. Fet amplifier with fixed bias.

table 3. Operating parameters of currently available power fets.

		Siliconix			Crystalonics			
	U183	U222	U244	U266	CP653	CP652	CP651	CP650
Gate-drain Gate-source	25	50	25	G-D 150	20	20	20	25
Maximum Voltage				G-S 20				
I_{DSS} (mA)	20	150	600	200	60	100	300	600
V_P	-8	-8	-8	-12	-5	-5	-5	-5
Maximum Drain Current (mA)	—	—	900	500	600	600	600	1200
Maximum Gate Current (mA)	10	25	—	—	—	—	—	—
Device Dissipation Ambient 25°C (W)	200	800	—	—	—	—	—	—
Device Dissipation case 25°C (W)	—	3	10	10	8	8	8	8
g_{fs}	5000	30000	150k	30000	6000	10000	10000	15000
C_{iss}(pF)	8	20	35	28	20	20	20	20
C_{rss} (pF)	4	4	15	16	—	—	—	—
Gate-source Breakdown Voltage	-25	-50	-25	-20	-20	-20	-20	-25
Power Output (W)	—	0.5	10	10	—	—	—	—

well as self-biased power oscillators. The high gate impedance has a minimum loading effect and oscillation starts easily in simple circuits. All the various crystal variations can be used — Pierce, Miller, Colpitts and modified Pierce — with a tuned output circuit. The simple Miller circuit works well. This can be a troublesome bipolar circuit at higher oscillator power levels. Furthermore, fets take off well in the push-pull circuits so popular in early vacuum-tube practice.

A practical push-pull circuit for 160-meters is shown in fig. 5. Power output of better than 1 watt can be obtained from this circuit using the component values shown. Inasmuch as cw operation is usually confined to a rather narrow span of frequencies on 160, the series variable in the gate circuit permits a helpful frequency spread. Low-cost U-183s and HEP-801s operate in the same circuit.



C1 1095 pF (three-gang broadcast variable)

L1 60 turns no. 26 enameled, closewound on 1¼" form, center-tapped. Leave space for L2 at center

L2 15 turns no. 26 enameled, closewound between two windings of L1

fig. 5. Push-pull fet crystal oscillator for 160 meters.

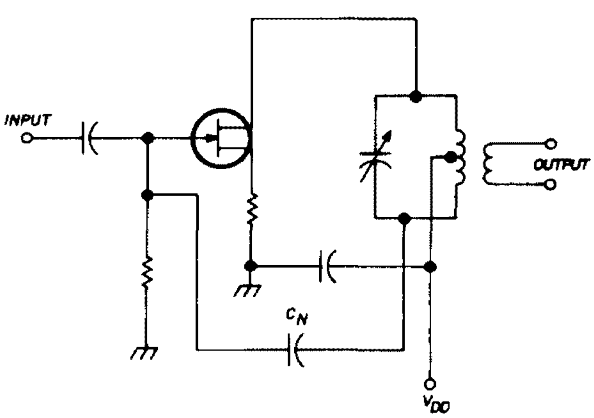


fig. 4. Signal biased fet amplifier.

wideband amplifier

Power fets do well as broadband amplifiers. The example in fig. 6 uses a Crystalonics CP651. Gain is 10 dB from 500 kHz to 40 MHz. Maximum input signal is 3 volts peak-to-peak across 50 ohms. A voltage gain of 10 dB builds this up to 9.5 volts across 50 ohms. Output power can be several hundred milliwatts.

fet harmonic generator

The high gate impedance of the field-effect transistor results in minimum crystal loading and an opportunity to emphasize harmonic output from a single stage. Such a circuit for obtaining good harmonic output was presented in *Electronics*,[2] **fig. 7.** The circuit reminds me of the old-fashioned tri-tet vacuum-tube crystal oscillator.

Fundamental oscillations are determined primarily by components in the gate-source circuit. The source circuit is tuned to a frequency somewhat lower than the crystal frequency (about 0.7 f_X). The resonant frequency must compensate for the influence of the gate-source capacitance of the fet.

The drain circuit must be set to the desired harmonic. Oldtimers will recall from past vacuum-tube experiments that this circuit produced a good odd-harmonic output and was often used as a tripler — and occasionally as a five-times crystal generator. Second and forth harmonic outputs are acceptable too.

Such a multiplier circuit is attractive for QRP operation because a 40-meter crystal will provide a strong 15-meter output. Forty-meter output can be obtained by tuning the drain circuit to the 7-MHz frequency and shorting out the source tank circuit. This circuit is also attractive for multiplying into the 2- and 6-meter bands.

vhf double sideband

Have we overlooked an economical means of transmitting sideband on the vhf bands? There are few amateurs with-

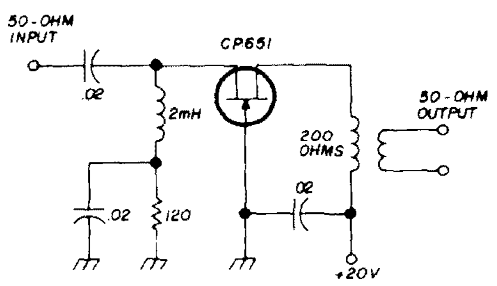


fig. 6. Wideband power amplifier. Response is flat from 500 kHz to 40 MHz.

appendix

The fundamental equation is

$$\frac{I_D}{I_{DSS}} = \left(\frac{V_P - V_{GS}}{V_P}\right)^2 \qquad (1)$$

In operation, the instantaneous total drain current is

$$i_D = I_{DSS}\left(1 - \frac{V_{GS}}{V_P}\right)^2 \qquad (2)$$

However, by expansion the fundamental component for a tuned rf amplifier is

$$i_D = 2I_{DSS}\left(\frac{V_{GS}}{V_P}\right)^2 \qquad (3)$$

The practical equations of the fet power amplifier are developed from **eq. 3.**

If the instantaneous i_D is to reach I_{DSS} on crest, (ie., $i_D = I_{DSS}$),

$$1 = 2\frac{v_{GS}}{V_P}$$

$$v_{GS} = \frac{V_P}{2}$$

This relation estimates a practical value of cutoff bias. For linear operation and a conduction angle of 180° the gate cutoff bias should be $V_P/2$. This ensures the signal swing can occur over the square-law region, and peak gate signal voltage reaches only zero gate bias.

Based on square-law transfer and $i_D(\text{peak}) = I_{DSS}$, the dc static drain current at zero signal level is

$$I_D = \frac{I_{DSS}}{4}$$

As the gate voltage swings between $V_P/2$ and 0, a load impedance of proper value produces maximum output and highest efficiency. The required load is

$$R_L = \frac{V_{DD} + (-V_P)}{I_{DSS} - I_D}$$

In class-C operation, a 120° conduction angle is obtained with twice cut-off bias

$$V_{cut\text{-}off} = (V_{P/2}) \times 2 = V_P$$

Bias is now at pinch-off value, and zero-signal static drain current is zero (I_D - 0) required load resistance becomes

$$R_L = \frac{V_{DD} + (-V_P)}{I_{DSS} - 0}$$

out the means of receiving high-frequency sideband signals; connect simple converters ahead of these receivers and you are blessed with an excellent vhf sideband receiver.

The transmit mode is more of a problem, requiring a transverter or a multiplier-mixer chain. You must start out with a sideband signal, and not everyone has this facility, particularly those who do all their operating on vhf.

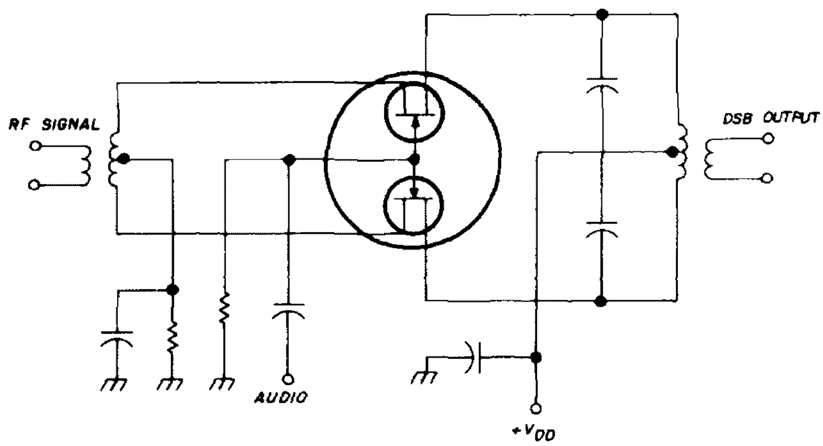


fig. 8. Dual-fet balanced modulator for double sideband generation. Capacitor C1 provides the carrier-balance adjustment.

The double-sideband technique provides an easy way of obtaining a sideband signal. In this mode of transmission the *carrier* is removed while both sidebands are transmitted. An ssb receiver is used for reception; either sideband can be received with appropriate setting of the receiver USB-LSB switch. A double-sideband signal isn't objectionable on the vhf

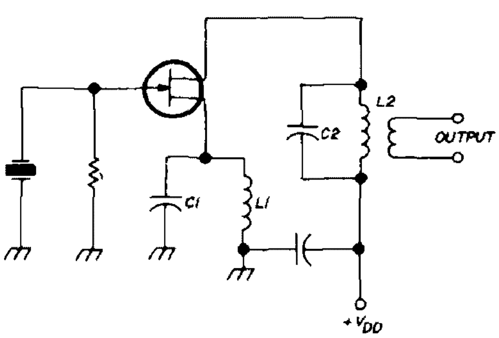


fig. 7. Harmonic crystal oscillator provides strong output at 2, 3, 4 and 5 times the crystal frequency. A HEP-802 works well in this circuit. C1 and L1 are tuned to the crystal frequency; C2 and L2 are tuned to the output harmonic.

bands because it occupies no more space than an a-m signal.

The dsb signal can be created right on the transmit frequency. Any number of techniques can be used, including high-powered vacuum tubes in a balanced push-pull circuit. Modulation can occur at low power level directly at the output of a crystal or vfo-controlled multiplier.

As a matter of fact, there will probably be a gradual displacement of the multiplier chains now so common, particularly on 2 and 6 meters. Stable solid-state vfos can be designed for operation directly on these bands.

There are any number of low cost crystal diodes that operate efficiently up to 500 MHz. These can be used in various balanced-modulator diode configurations. Field-effect transistors and integrated circuits are also available for this use. The dual-fet, **fig. 8,** has good possibility. The circuit is so simple and the components so few that good performance is almost a certainty. If you are a purist, this configuration can be connected in a double-balanced modulator as an ideal way of generating a phasing-type single-sideband signal. We'll dig into this one, too, a little later.

references

1. Siliconix applications note, "High–Frequency Power Field-Effect Transistors."
2. Fred B. Cupp, "Fet Harmonic Generator," *Electronics,* 18 September, 1967.

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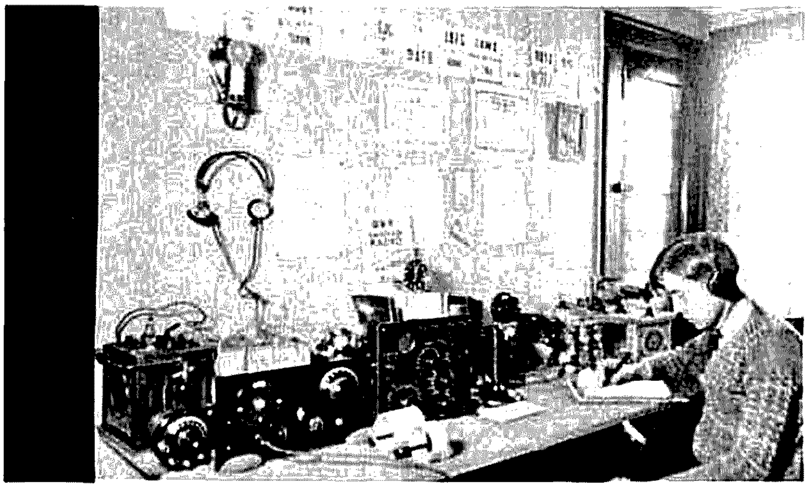


those were the days

An invitation to share some happy memories with an oldtime radioman

John D. Davies, K4NW, Box 188, Rockledge, Florida 32955

From time to time *ham radio* has presented articles on the early days of wireless and the work of some of the pioneers who have contributed to the art of radio communication. While the space devoted to these pieces might well have been used for more items on today's technical problems and their solutions, we at *ham radio* feel that an occasional digression into radio's past helps preserve the editorial balance we're trying to achieve – in short, a magazine with "something for everyone."

This article is dedicated to the old timers – a very substantial portion of the ham fraternity. Many are still active, even after fifty years or more. What's more important, many old timers have kept abreast of the rapidly changing technology in electronics and have made significant contributions to the state of the art. And that's a healthy sign.

For interested readers, a bibliography is included at the end of K4NW's old-time radio story. These are articles we've published over the years by another old timer who has kept up with the action – Ed Marriner, W6BLZ. **editor**.

Radio communication from 1920 through the early Thirties held a peculiar fascination for the public, perhaps because people then were less sophisticated than the present generation. The air of mystery surrounding early radio seemed to motivate even those with no knowledge of electricity into trying their hand at constructing a radio set. Some were content to replace their primitive sets with manufactured models as they became more numerous and as prices became more reasonable. Others, unable to resist the urge to experiment, continued to learn, to innovate, to improve; and so ham radio began to flourish.

Come with me for a few nostalgic moments and we'll look at some of the early hardware and ideas that formed the basis of today's electronics.

early equipment

If you were a teenager around the first half of the Twenties, surely you'll remember the Reinartz circuit; three-circuit

tuners; honeycomb and bank-wound coils; Daven resistors; Radiola, Grebe, and Paragon receivers; and a host of other components and equipment. Remember the smell of shellac on homewound coils; the sound of chemical rectifiers bubbling inside Mason jars; VT1 and VT2 tubes; the whine of a correctly adjusted synchronous spark gap? – Ah, sweet memories!

Me? I started my radio career with a crystal set. The inductor was wound on a Quaker Oats box using no. 22 dcc wire. Then the insulation was carefully scraped off, and taps were soldered onto the wire for a switch. I didn't use a variable condenser – just switch points. The rest of the circuit consisted of a galena-and-catwhisker detector, a "phone condenser" of tinfoil and waxed paper sandwiched together, and earphones. This was around 1921 and '22. Soon thereafter this deluxe layout was followed by an "Arlington Loose Coupler" and a "Fada Crystal Detector."

publications

Early newspapers featured a "radio section" offering the latest do-it-yourself

The operating corner of W2BKD's shack.

project – a reflex, neutrodyne, or similar receiver circuit (they all worked about the same) – together with the latest version of the popular "three-circuit tuner." All were accompanied by more or less authentic advice, depending on who happened to be the current "radio editor."

Radio magazines of the day were sometimes a wonder to behold. Those wishing to build equipment from the published circuits followed instructions to the letter. If the instructions called for double-cotton-covered wire, for example, you'd never dream of using single-cotton-covered wire – it was all that mysterious.

Many early radio magazines carried Q & A columns for aspiring radio constructors. Some of the questions and answers were interesting. A typical query, appearing in a 1923 edition of a popular mag, went like this:

"Q. What is the advantage of connecting the tube-filament rheostat in the negative wire of the 'A' battery circuit?"

The editor, who was somewhat of a diplomat and a little unsure of the answer himself, replied:

"A. Better results will be obtained if the rheostat is placed in the negative lead of the 'A' battery. This is particularly true of the UV-201A tube . . ."

A circuit then followed showing an elaborate switching arrangement that allowed the filament rheostat to be transferred between positive and negative leads of the filament supply. (The editor took no chances.)

The same blind faith went for antennas. If the article or book you were following said to use no. 14 7-strand copper, you wouldn't dare use no. 12 solid. Arguments were rife on the merits of four versus six parallel wires between the antenna spreaders. Some swore by cage antennas. After all, if your neighbor heard PWX (Havana) with a cage, what more proof did you need? Before long, though, some DX hound who finally at 3 a. m. snagged KFI in Los Angeles would claim that a 150-foot-long wire was superior to all spreader-type aerials. Who were you to question such a feat? Down

with the cage; up with the long wire. And so it went.

broadcast stations

In the early days, many experimenters were interested in receiving the radio broadcast stations that began to appear all over the world. There was, of course, KDKA in Pittsburgh; WOC in Davenport; Milton J. Cross in Aeolian Hall at WJZ. Then there were CZE, Mexico City; 2LO in London with Big Ben (live) tolling midnight to the tune of background music furnished by a dance band at Ciro's.

The "International Tests," held once a year, were highlights for the DX listener. American broadcast stations would shut down for a time so we could listen for European stations; while another night the Latin Americans would come through.

Then there were the corny jokes of the DX hounds: "I opened the window last night and got Chile;" "My wife was in the kitchen and I heard China;" "Are you Hungary? I'll Russia to the table and Fiji" – to name a few. Yes, those were the days!

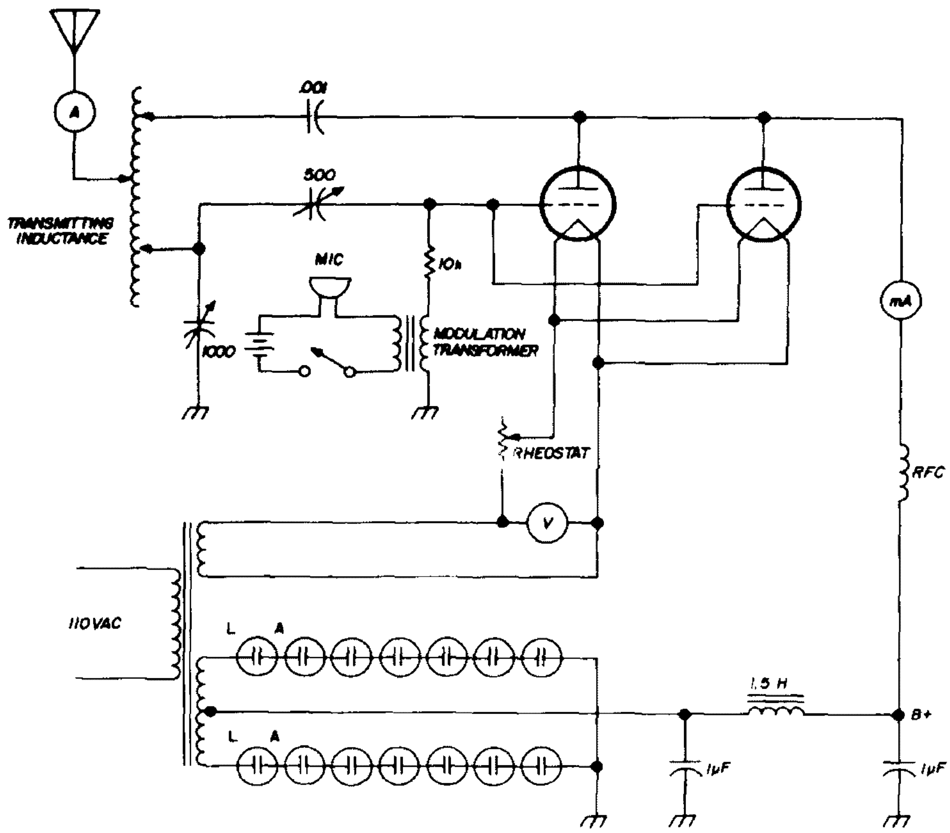


fig. 1. A 10-watt cw and phone transmitter popular during the early Twenties. Power was supplied to a Colpitts circuit by electrolytic rectifiers. "Very good results" were claimed for this "Hook-Up."

spark transmitters

I was a bit too young to be identified closely with spark, except for experimenting with a certain portion of Henry Ford's ignition system – but I knew spark was around. At frequent intervals a foreign ship would drop anchor in the Hudson and open up with what I understood to be, in ham terminology, an "Old Betsy." The operator would tear the air apart for a while then move on, to be replaced by another vessel with, perhaps, an old Federal arc transmitter whose sound was pure music compared to the spark transmitter.

the "short" waves

One day word got around that amateur radio operators had "spanned

the Atlantic!" I didn't have a ticket then. I was still "bootlegging" across town with a Ford spark coil and wondering whether a hot pink spark was better than a cool blue one. But I started winding coils with fewer turns, and to this day I've never left that wonderful land "below 200 meters."

When I finally landed a ticket I tried to get my ac-powered oscillator somewhere near the 40-meter band. I listened

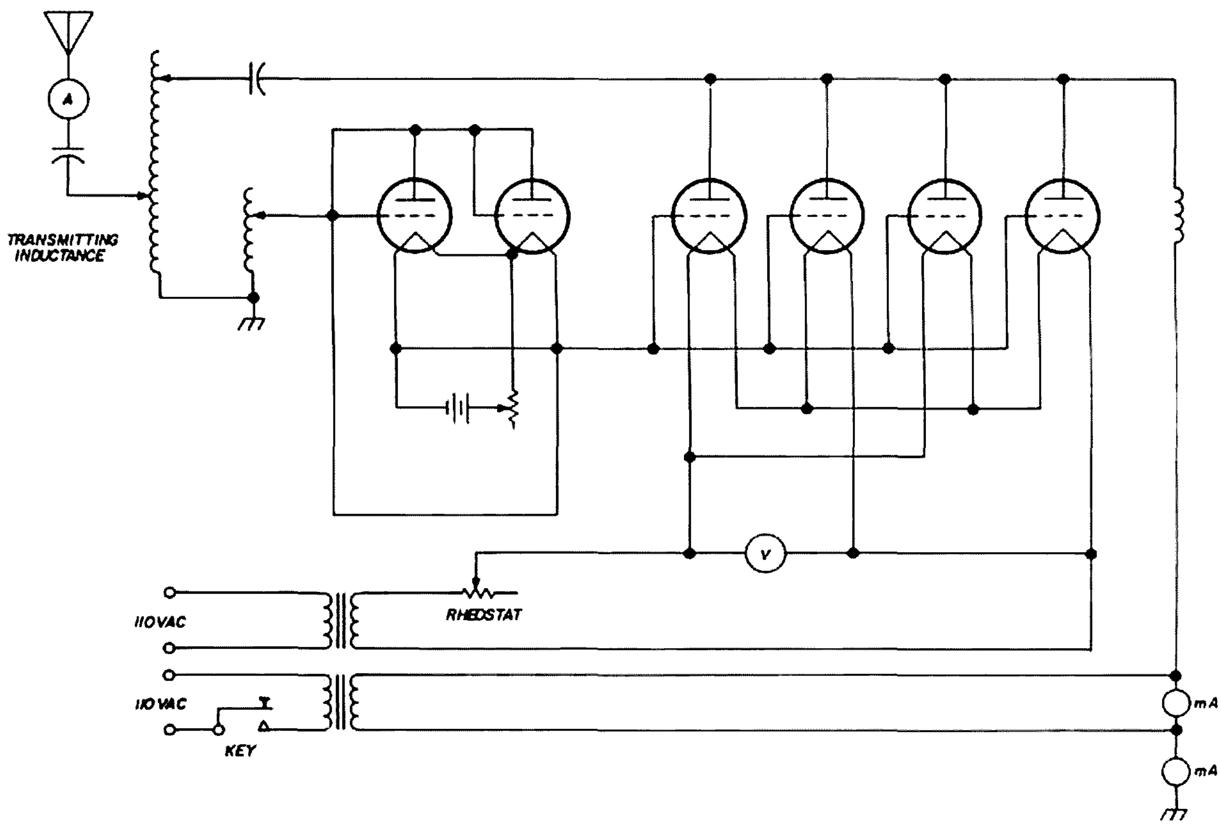
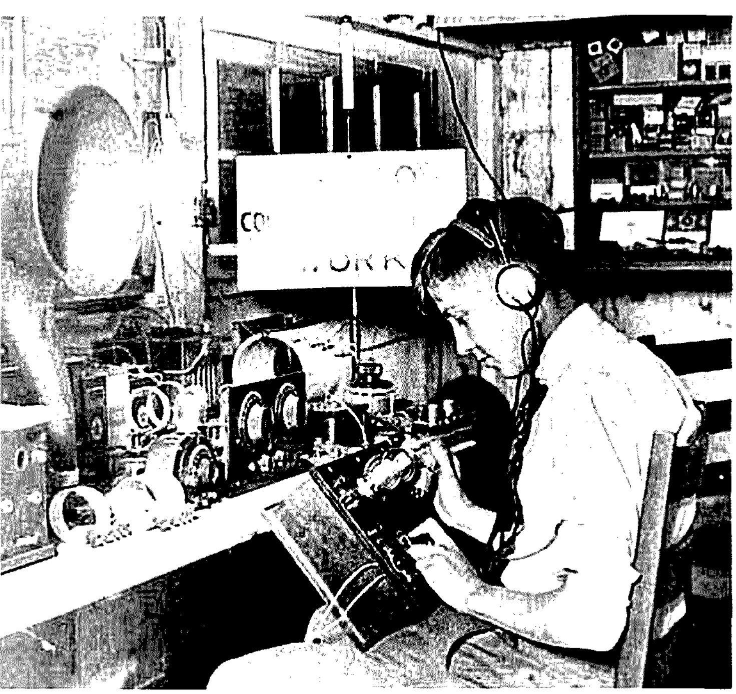


fig. 2. French station 8AB's transmitter, heard often in the U. S. during the Trans-Atlantic tests. Rig ran 1 kW input, using French 250-watters. Two keys were necessary "to allow the hot one to cool while the second was in operation."

Troubleshooting a "tough dog" (c. 1930).



for a CQ that first day, and finally there it was: NU1AD (NU was shortly replaced with W in U. S. amateur call signs).

I timidly clicked out NU1AD's call. In those days you didn't listen on your frequency; you "combed the band" for replies. Lo and behold: NU1AD answered! When I told him (C. S. Doe, Bellow Falls, Vermont) that he was my first contact, he calmly recounted how he'd worked across Boston harbor with a spark set in 1916 for *his* first radio contact. Nowadays I tell the young novices how I worked all the way from New York to Vermont in 1928 with 5 watts.

transmitters

The rig responsible for my first contact used a single 210 (UV-210, rated at

7-1/2 watts) in a Hartley circuit, fed by all the ac I could muster. Today it would make a first-rate jammer. The antenna was a voltage-fed Hertz, very popular at the time along with the Zepp.

Other equipment included an ancient RCA transmitter "appropriated" from a tugboat somewhere in NY harbor. This rig used four UV-202s and four Kenotron rectifiers. For "radiophone" you modulated two of the 202s paralleled in an oscillator, with the other pair paralleled as modulators. Each tube was rated at 5 watts. In the "telegraph" mode, all four 202s were paralleled for a full 20 watts out! But I never could work anybody with that old thing.

Under the bench was a collection of "slop jars," which bubbled and boiled so furiously when I pressed the key (despite the oil on the troubled waters) that I had long before given up trying to rectify the 110-volt ac line. A raw ac note was acceptable in those days, anyway. Plenty of rich guys used 852s (75 whole watts!) fired with straight ac, so why shouldn't I use ac?

"high" power and DX

As time went on I coaxed more and more out of the old 210 'til I had her running at around 15 or 20 watts. The plate would go from red to yellow if you made your dashes too long; but everybody knew this and didn't get unduly alarmed, unless the end of the dash slithered off so far you had to retune the receiver. Though one time when I was playing with an old WE 216A I did put a neat little hole in its corrugated plate.

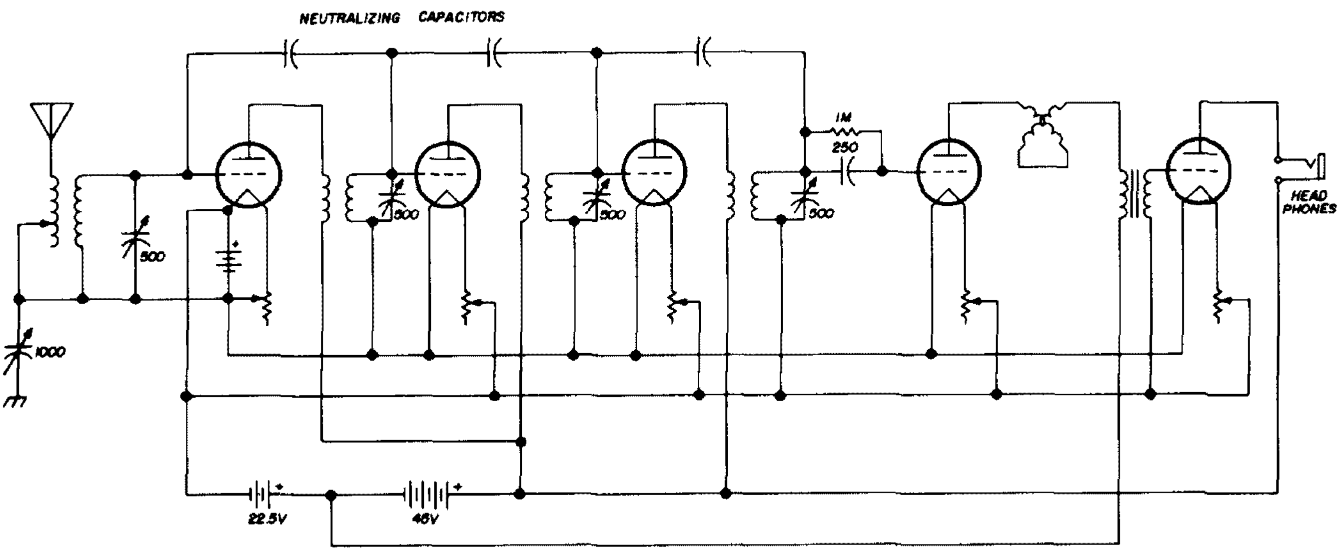


fig. 3. The Neutrodyne receiver, using 3 tuned rf stages. It was a bear to tune, but really brought in the DX. Note the neutralizing scheme — capacitor values were determined by cut and try.

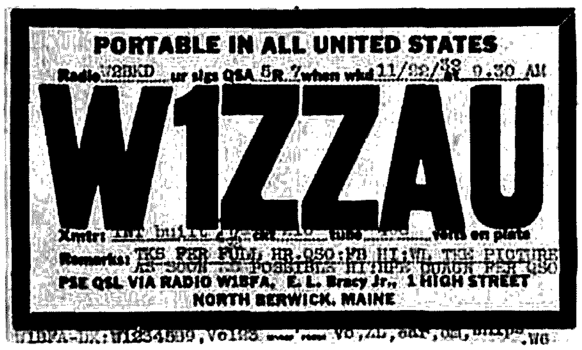
And I got that voltage-fed Hertz a little higher up by using a bow and arrow to shoot a thread over a high branch of an old locust tree. With the thread I pulled up a string, followed by a rope, followed by the antenna itself. That helped.

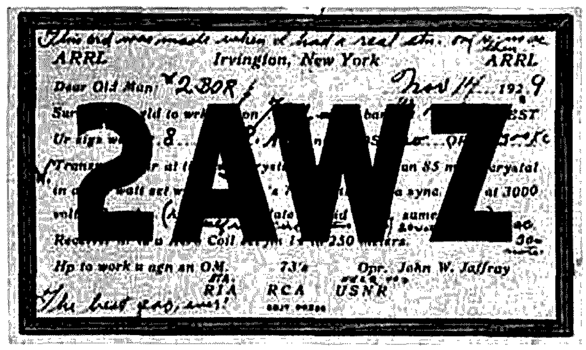
By 1929 I had actually worked a VK. They had changed from OZ (for Oceania, Australia) to VK along with our switch to W from NU (for North America, U. S.). And, by the way, that Aussie was using an old 171A in his transmitter! In 1929, too, we had "international" ham bands. Prior to that you'd often find the foreigners in separate bands of their own.

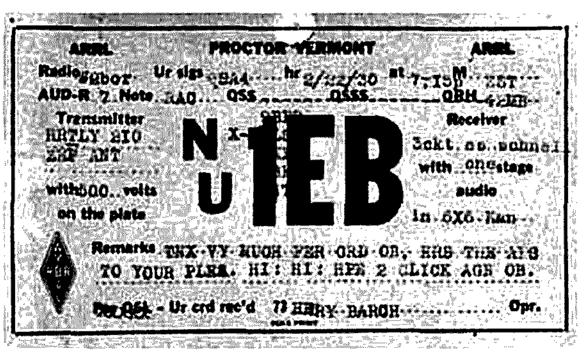
welcome to the shack

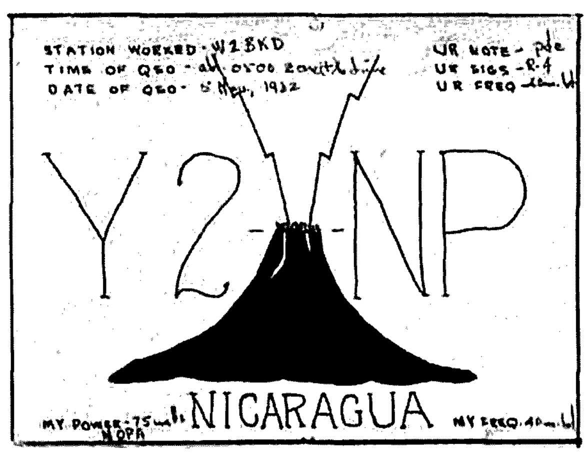
In front of me are some beaten-up old photos of the shack, taken at various times during the Twenties. Over there is a helix, an 8-inch-diameter coil of flat nickel-plated ribbon with one edge buried in a

wood base. It came out of that old RCA tugboat transmitter. Yonder is the familiar old upright telephone mike. The plug-in coils of the receiver stand out clearly. Down at the end of the bench is an old long-wave receiver, an IP-501.









QSL cards typical of the early radio years.

Hanging on the wall amidst the ancient OSLs is a picture of Graham MacNamee, idol of announcer-worshippers, and certainly a far cry from today's "DJs." Back of the 3-circuit regen with the plug-in coils is arrayed a host of 45-volt B batteries and a 6-volt battery, along with the familiar old 4½-volt C battery.

and, of course, QSL cards

QSL bureaus were unheard of. Everything was "direct." A lot of OSLs were homebrew like the rigs themselves, especially during the Depression. But we relished 'em. Today the choice ones hang in my shack, yellowed with age, but protected by those fancy new cellophane holders.

You can see W1ZZAU (portables got calls like that!), old CM1ML, 2AWZ (Jaffray), NC3AY, 10016, U1QA, and many others.

now, a parting word

Because I still grow nostaligic at the smell of shellac and soldering flux, most of my gear today is still homebrew. But I'll admit it's mighty hard to equal the performance of today's ready-made equipment. Guess the best we can do is compromise by building a little and buying a little.

Still, it's nice to dream, isn't it – of UV-202s; WD11s, the beautiful music of an icw signal; KDKA; that first contact; that first overseas DX; I could go on for hours.

Those were the days, my friend.

bibliography

1. E. Marriner, W6BLZ, "Early Wireless Stations," *ham radio,* October, 1968, p. 64.
2. E. Marriner, W6BLZ, "Wireless Point Loma," *ham radio,* April, 1969, p. 54.
3. E. Marriner, W6BLZ, "Catalina Wireless 1902," *ham radio,* April, 1970, p. 32.

ham radio

television interference: an effective remedy

TVI can exist even with today's modern ham gear — here's one solution to the problem

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I've been an active ham for 36 years. During this time I've built many low-power rigs. I guess I enjoy working *on* a transmitter as much as working *with* it. I like all bands, so I'm partial to a long-wire antenna. Recently I began using an indoor wire. This antenna provides excellent DX reports, but it doesn't discriminate against TVI.

A few months ago I retired from home-brew rigs and purchased a Swan 350C, a transceiver with many good features and a lot of power for the money. I soon discovered that when operating on 20 meters, my favorite band, the Swan caused interference with my wife's TV programs, particularly on channel 2. (She watches a portable in the next room, about 12 feet from the rig and antenna.) On the other hand, 15 meters was perfectly clean on all low TV channels, but TVI showed up to a lesser degree on channel 9 and higher.

After discussing the problem with fellow amateurs, I realized that something had to give, and it turned out to be me. I would have to sacrifice some power. Fortunately this was not difficult to accomplish, and the difference in signal reports was insignificant.

the cure

It's quite possible that the modifications I made will apply to equipment other than Swan. Here's what I did, in the order of effectiveness, to cure the TVI problem:

1. Input to the driver stage, a 6GK6, was reduced, which decreased harmonic output.

2. Input to the final, which uses a pair of 6LQ6/6JE6 tubes, was reduced from 360 to 200 watts.*

3. Capacitors (1 kV disc type, 0.001

*This is a decrease of about 2.5 dB, which is barely perceptible under *ideal* conditions. It's unlikely that the receiving station could tell the difference between 360 and 200 watts under actual band conditions. *editor.*

μF) were placed across filament, high-voltage, medium voltage, and bias terminals — right at the Jones plug. (These are leads between the transceiver and power supply.) Also I shortened the unshielded cable between the rig and power supply, which reduced radiation from the cable.

other changes

The Swan has three supplies: 850, 275, and 215 volts. This last voltage is obtained through a dropping resistor. The Swan schematic shows that the 850 volts is the sum of two supplies in series: 275 and 575 volts. I separated these supplies and used them independently. The 575 volts was applied to the final plates (instead of 850 volts); then I added a 150-volt zener, fed from the 215-volt

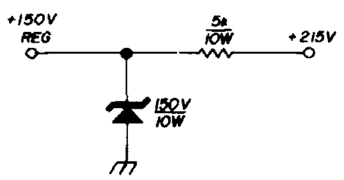

fig. 1. Zener diode added to the Swan 215-volt supply for final and driver screens to reduce input power for TVI reduction.

supply (**fig. 1**). The 150-volt supply is now used for the final and driver screens, The "new" and "old" voltages are listed in **table 1**.

commercial data

Lest these changes seem drastic, I'll reference an article appearing in *Sylvania News.*[1] This article discusses several tubes, including the 6JE6. The following optimum values are suggested for the 6JE6:

plate voltage	500
screen voltage	125
bias voltage	-85
plate current	222 mA

These values are recommended in the Sylvania article for class-C cw operation. For class AB1, the same maximum plate

table 1. Voltages for the Swan 350C before and after the TVI modifications.

	driver		final	
	plate	screen	plate	screen
new	270	150	525	150
old	270	270	850	210

and screen voltages are recommended, plus

idle plate current	40 mA
bias voltage	-44 Vdc

Note that these are still slightly below those for the "new" values given in **table 1**.

results

When the changes described above were incorporated, TVI was greatly reduced. I can now monitor a portable TV receiver with its single collapsible rod antenna about two feet from my indoor transmitting antenna. On channel 2, barely visible lines appear; but previously the picture was completely wiped out! With the TV set in the next room, no lines are visible at all. It's almost impossible to tell when I'm on the air. The loss in transmitter power is less than 2:1, so the sacrifice is small.

further power reduction

If you wish to reduce transmitter power still further, disconnect the input capacitor to the filter in the 275-volt supply. The voltage will be decreased to about 200 volts. Input to the driver will be much lower, reducing harmonic output still further; and the dc input to the final will be about 125 watts. The lower voltage will also be applied to the tubes in the receiver section; however, I didn't notice much loss in sensitivity. If you have a really severe TVI problem, this certainly should do the trick.

reference

1. W. D. Murphy, "Horizontal Deflection Tubes as Power Amplifiers," *Sylvania News,* Vol. 31, No. 4, Winter, 1964. Sylvania Electronic Components, Waltham, Massachusetts 02154 ($1.00 per bound volume).

ham radio

charge flow in semiconductors

This will help you understand the principles of transistor theory

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Electron flow in vacuum tubes is not hard to understand, but many hams find it difficult to shift from tube to transistor theory. Indeed, a physicist could spend most of his life studying transistor theory, only to find that the more he knows the more there is to learn. Fortunately, it's possible to simplify transistor theory so that almost all of the happenings in the transistor can be explained in simple terms.

This article is written for the ham who wishes to understand the basic interactions that take place in the six-thousand-plus different types of transistors on the market today.

Whether a material is a conductor or a dielectric depends on the quantity of electrons that are rigidly bound within the material's atomic structure. If the material has an excess of loosely held electrons, the electrons can be easily disturbed by an external influence such as an electrical force. The material is then said to be a conductor of electricity. If, on the other hand, the material has few loosely held (free) electrons, the material has a high resistivity. The electrons are not free to distribute themselves readily, and the material is called a dielectric or insulator.

Transistors are made from crystalline substances called semiconductors. The atomic structure of these substances places them about midway in the resistivity scale between conductors and dielectrics. Modern transistor theory is based on how these crystal substances behave when their atomic binding forces are disturbed in a controlled manner, as by adding impurities. This is the basis of the discussion to follow.

atomic structure

According to the Danish physicist, Neils Bohr, the atom has a positively charged nucleus surrounded by orbiting electrons. Each orbit has a distinct number of electrons, and each electron has a distinct energy level. The electrons in the outermost orbit have the highest energy level (highest momentum). They are also farthest from the binding force of the atom's nucleus, hence they can more

easily break free from the parent atom.

The crystal substances used in today's transistors and diodes are germanium and silicon. The outer orbits of these materials have four electrons. The balance between nucleus and electrons prevents the material, in its natural and undisturbed state, from acquiring any more electrons. Thus in an ideal crystal, where the electrons are bound firmly within the structure, no conduction occurs. The material is, theoretically, a perfect dielectric. A two-dimensional representation of the structure is shown in fig. 1. Here, each atom shares its four electrons with its neighbors.

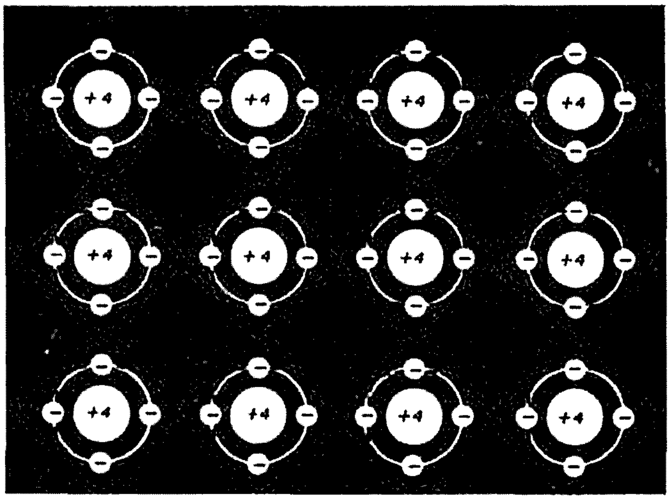


fig. 1. Atoms bound together by sharing electrons in a pure silicon or germanium crystal.

Free-moving electrons are necessary for conduction in any substance. If the electrons are held tightly in the structure, as in fig. 1, no conduction occurs. However, if enough energy is available, an electron can be freed from the bond.

Heat is one form of energy that can free electrons. At room temperature, there is enough energy to free some electrons and provide conduction. The more heat put into the crystal, the more free electrons, and the higher the conduction.

holes

Every discussion on transistor principles mentions holes and hole conduction. This is perhaps one of the most difficult concepts to grasp. Whether you're talking about transistors or doughnuts (they're both made of matter), a hole is a very real thing and can't be ignored. I'll try to explain why.

A crystal of pure germanium or silicon is a dielectric. No conduction occurs unless an electron breaks free from the bond existing between neighboring atoms. If energy is applied, however, an electron is free to move through the structure. Conduction is then said to exist.

Electrons possess a negative charge; hence they are called negative-charge carriers. When an electron is freed from its bond, a hole is said to exist. What is a hole? It's either something or nothing, depending on your viewpoint. In electron physics it's a very real something.

When an electron is knocked out of its orderly path in a substance, the space it formerly occupied *must* be accounted for. The space vacated by an electron is called a positive-charge carrier. It can be shown mathematically that this carrier moves from place-to-place within the substance. Holes can carry a charge, because they cause the structure to become positive *without* the electron. The idea of holes and electrons is shown in fig. 2. It's important to understand that the hole *must* exist, because making a

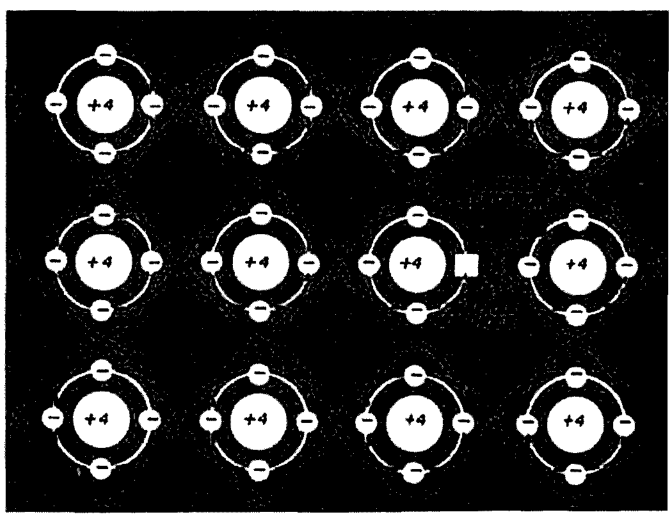


fig. 2. Free hole and free electron created by thermal energy breaking the bonds.

hole in nothing would be meaningless. It may be difficult to visualize something that is really nothing, but the hole does *do* something.

The electron, unlike the hole, can exist by itself and doesn't have to hop from one atom to another. It can move more independently through the crystal, and it happens that the electron moves about twice as fast as the hole.

Fig. 3 demonstrates the idea of conduction. We use a piece of silicon with a battery connected across it. The battery is simply a pump that pumps electrons from its positive end out its negative end. Half as many holes flow in the opposite direction. The holes are pumped out of the positive side of the battery and into the negative side. Whenever a bond in the atom is broken to create an electron, a hole is formed, so an equal number of holes and electrons exists. However, since the electrons move twice as fast as the holes, there will be *twice* as much

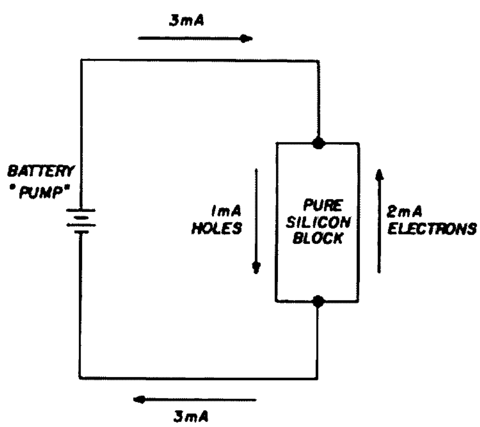


fig. 3. Charge flow. Current is the sum of the holes and electrons and is from positive to negative.

electron current as hole current. If we assume a two-milliampere electron flow, then there will be a one-milliampere hole flow in the opposite direction.

Which way does the current flow? This is an ambiguous question, because current actually flows in *both* directions. If only the word *current* is used, it usually means a charge going from positive to negative in the circuit. In most textbooks (and in this article), current always means the *total* charge flow in the positive-to-negative direction. If we have two milliamperes of electron charge flowing from negative to positive, and one milliampere of hole charge flowing from positive to negative, *three* milliamperes of current will flow from positive to negative.

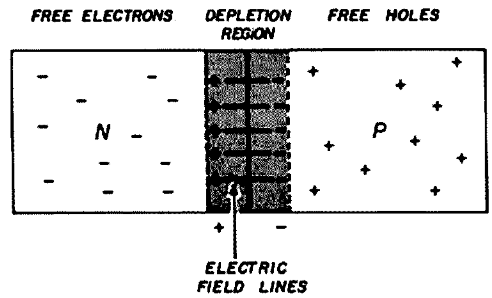


fig. 4. Depletion region in p-n junction.

doping

As stated earlier, the semiconductor material is not a very good conductor, because not many charge carriers are freed from the bond between atoms. However, it's possible to create charge carriers in larger numbers by a process called *doping.*

Suppose we add some atoms with either three or five electrons in their outer orbits to the pure germanium or silicon. Several elements possess this structure. Arsenic and indium are the most popular materials for doping. The foreign atoms in these materials will mix into the crystal and form bonds with its pure semiconductor material. However, if only three electrons exist in the outer orbit of a doping atom, a hole will exist in the structure. Just the opposite happens with five electrons instead of four. In this case there's an extra electron. This is how *free-charge carriers* are created.

If we dope a piece of silicon with atoms that have only three electrons in their outer orbit, the majority of the charge carriers will be *positive holes.* This is called P-type semiconductor material. If the doping atoms have five electrons in the outer orbit, then we have electrons as

the majority-charge carriers. This is called N-type material.

the p-n junction

Suppose a block of N material is attached to one of P material. The electrons are still free, bouncing around with thermal energy in the N-type material, and the holes bounce around in

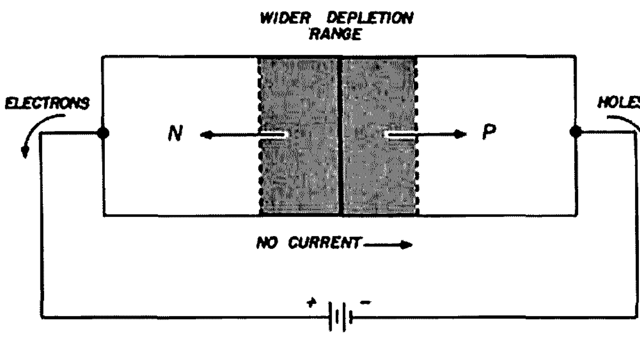


fig. 5. Reverse bias. The depletion region is widened so that no current flows.

the P-type material. The charge is still balanced, and the holes and electrons move randomly. Suddenly an electron strays across the P-N junction. It is instantly captured by a hole and held tightly in the bond. An electron has left the N side, so the charge balance is upset, and the N side is positive by one charge. For every charge carrier that wanders from one side to the other, a line of electric flux is formed. Charges continue to bounce across the junction, driven randomly by heat energy, building up an electric field at the junction. The positive polarity of the field is on the N side, and the negative polarity is on the P side. Soon the field builds up to a point where the charge of the field overcomes the thermal energy. Since there is a large negative charge on the P side, electrons are repelled from crossing the depletion region to the P side. **Fig. 4** shows this situation. The charged area is called the depletion region, because all electrons and holes have combined. All the moving charges have depleted and are in equilibrium. After studying fig. **4**, this section should be reviewed, because this is an important concept.

the diode

Now suppose we attach a battery across the P-N junction. The positive side of the battery is connected to the N-type material; the negative side to the P-type material. The battery pumps holes into the negative side and pumps electrons into the positive side. At each end of the semiconductor material, charge carriers of one kind are pumped out and replaced by those of the opposite kind, which recombine and become an inert part of the structure. This is illustrated in fig. **5.** As a result, more charge carriers are depleted, and the depletion region becomes wider. This makes the barrier higher, which prevents carriers from crossing by heat energy. Since no charge carriers can cross, no current flows.

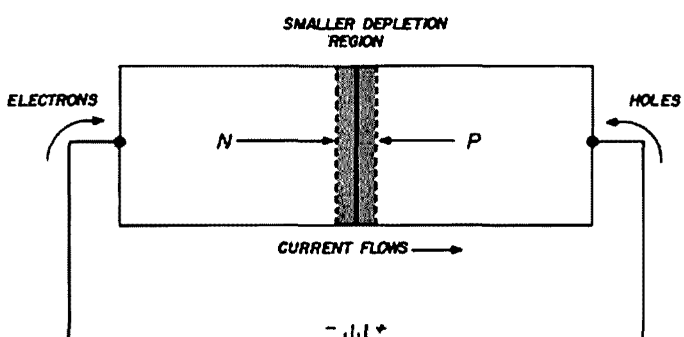


fig. 6. Forward bias. The depletion region is reduced, allowing current to flow.

Now suppose the battery polarity is reversed, as in fig. **6.** The battery pumps electrons into the N material and pumps holes into the P material. This increases the charge carrier concentration. More

holes are on the P side, and more electrons are on the N side. Thus the depletion region is smaller. When the barrier is small enough, the charge carriers will cross. Carefully note that conduction *is not* a result of carriers being attracted by the battery. Conduction results from the random movement of carriers that have increased in quantity. Conduction also results because of the lowered barrier.

So now we have a diode. But the real magic is yet to come.

the transistor

Suppose we add another N-type material to the P side of the P-N junction. This is then an NPN arrangement, as shown in fig. 7. The P part is usually made very thin; 0.001 inch or less. Since this is a transistor, the N-type material is called the emitter, the P-type material is the base, and the N-type material is the collector.

Batteries and ammeters have been connected to the transistor (fig. 8). The batteries act in the same way as in the diode discussion, i. e., pumping holes out of the positive side and pumping electrons out of the negative terminal. The batteries can also be thought of as pumping electrons into the positive side and pumping holes into the negative side. It will be necessary now to call upon some circuit theory to explain the current flow in fig. 8.

Kirchoff's law states that the sum of the currents in a closed loop must be equal to zero. This means that the sum of the currents coming out of a branch must equal the sum of the currents going in.

Consider the base-emitter junction. Note that this junction is forward biased, and the depletion region is lowered. The electrons in the emitter can now cross the low depletion region as can the holes in the base. The electrons move about *twice* as fast as the holes. With this thought in mind, suppose that two milliamperes of electron current flow from emitter to base, and one milliampere of hole charge flows from base to emitter. Since electrons and holes move in opposite directions, they will *add* in the emitter circuit, so that three milliamperes will flow in the emitter circuit. The I_E meter will read 3 Ma. The electrons flowing from emitter to base have tremendous velocity, and most cross the base into the collector. About one percent fall into holes in the base; but for practical purposes, we'll assume they all enter the collector.

The battery is connected so that the

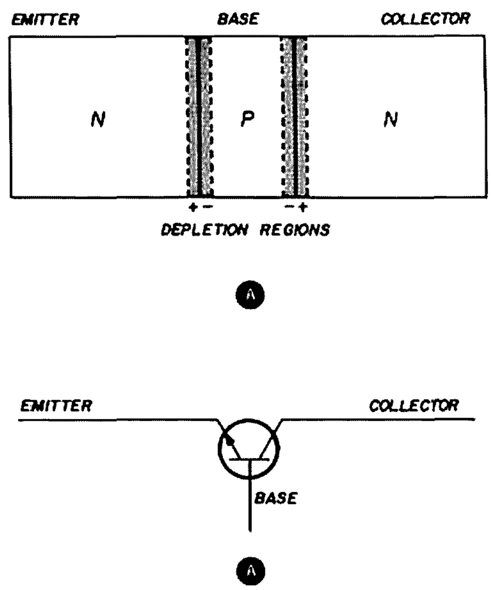


fig. 7. The npn transistor. (A) shows the junction of two materials having an excess of free electrons (n type) and an excess of positive charge carriers, or holes (p type). The schematic symbol is shown in (B).

collector-base junction is reverse biased. Therefore, electrons are pumped *out* of the collector. Before the emitter current becomes effective, no current can flow because of the reverse bias. But the two milliampers flowing into the collector are an *excess.* These electrons are pumped out of the collector, causing two milliamperes of current to flow. The meter, I_C, will read 2 Ma.

To find the base current, I_B, we must consider the connection between the batteries. Three milliamperes are pumped out of the emitter by the emitter battery, B_E. Since two milliamperes are pumped out of the collector battery, B_C, two of the three milliamperes from B_E *must* flow through B_C. One milliampere has to

go somewhere, so it must go into the base. Recalling that one milliampere of hole charge flows from base to emitter, the circuit checks out.

amplification

If the emitter-base voltage of fig. **8** is varied, thus varying base current, then the current from the emitter varies. This varies the collector current. A small change in base current causes a much

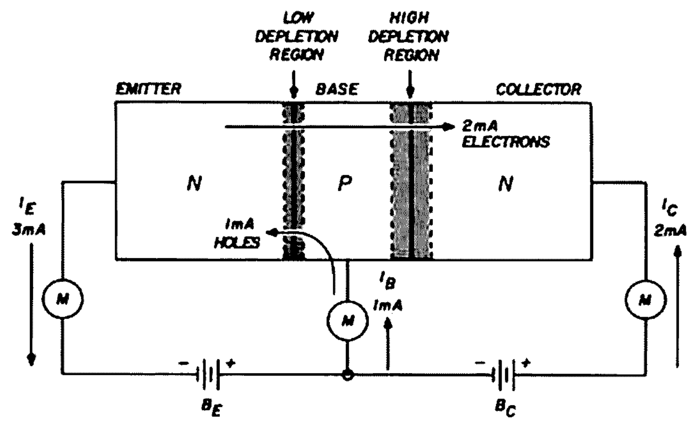


fig. 8. Npn transistor with batteries showing current flow. The collector current is larger than the base current, causing amplification.

larger change in collector current. Thus we have gain, or amplification. The dc current gain of a transistor, called beta (β), is the ratio of collector-to-base current:

$$\beta = \frac{I_C}{I_B} = \frac{2}{1} = 2$$

If one milliampere flows into the base, two milliamperes will flow from the collector.

doping levels

A gain of two is pretty poor by today's standards. The problem is due to the large base current. If the base current could be reduced, while retaining the same collector current, the gain could be increased.

One way to reduce base current is to reduce the flow of hole charges from base to emitter, because this is the only significant charge flow from the base. The hole charge flow can be reduced by reducing the amount of base doping, which impedes the number of holes in the base. If base doping is reduced by one hundred times, the hole current from the base will be reduced to 0.01 milliampere. If we make this modification to fig. **8**, the base current will be 0.01 Ma. Since two milliamperes flow from the emitter, two milliamperes will flow into the collector. The dc current gain is then

$$\frac{I_C}{I_B} = \frac{2}{0.01} = 200 \text{ mA}$$

In modern transistors, the emitter is doped from 500 to 1000 times more heavily than the base. Several factors limit the amount of doping. If the base is doped too lightly, the device becomes an insulator, and no current flows. It's desirable to dope the collector as heavily as possible to reduce its series resistance. However, this reduces the breakdown voltage. If the base is made thinner, gain is increased, because there is less chance for an electron to cross the base and be caught by a hole. If the base is *too* thin, the two depletion regions on either side will join when high voltage is applied. Then the transistor will short circuit, causing destructive current flow.

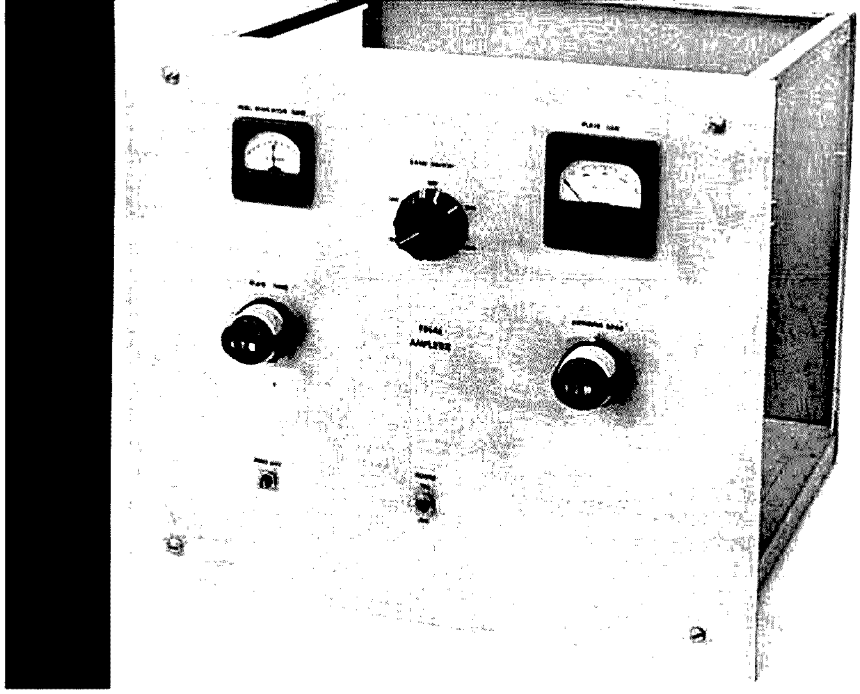
conclusion

I have made omissions and approximations to keep the discussion simple. Many more facts about semiconductors are available in hundreds of textbooks. The main purpose of this article is to present the basic concepts of these marvelous devices in an easy-to-understand manner. I hope I've succeded.

bibliography

1. "G. E. Transistor Manual," J. F. Cleary, Editor. General Electric Company, Electronics Park, Syracuse, N. Y.
2. Keith Henney, "Radio Engineering Handbook," McGraw-Hill Book Co., New York.
3. "RCA Transistor Manual," RCA Semiconductor and Materials Division, Somerville, N. J.
4. F. E. Terman, "Electronic and Radio Engineering," McGraw-Hill Book Co., New York.

ham radio



high-power linear
for
80-10 meters

The Eimac 4CX1500B
in a circuit
with novel tuning
and bias-control features

Irwin R. Wolfe, W6HHN, 3467 Rambow Drive, Palo Alto, California 94306

The 4CX1500B ceramic tetrode is used in many commercial high-frequency transmitters, but this tube seems to have been overlooked by amateurs. It has some features that should appeal to the "home-brew" ham, such as:

1. Lower grid-drive requirement than for 1-kW triodes.

2. No neutralization.

3. Low intermodulation distortion in linear-amplifier service.

4. High plate dissipation, which means longer tube life and less chance of tube destruction when operated in an off-resonant condition.

As with other tetrodes in this class, certain design criteria and precautions must be observed with the 4CX1500B to obtain satisfactory operation within published ratings. Some of this tube data is presented at the end of the article. It is recommended that those wishing to build the amplifier described here become familiar with all the characteristics of the tube (reference 2).

amplifier circuit

Standard components are used throughout. Although vacuum variable capacitors are shown in the schematic (fig. 1), air-dielectric variables can be used, and a multi-turn dial won't be necessary (see photo).

The amplifier employs an untuned input circuit and a conventional pi-network output circuit. The rf drive signal is terminated in a 50-ohm noninductive resistor through which grid bias is supplied to the tube. To improve the input-impedance match at the three higher-

frequency amateur bands, three untuned grid coils are used to approximately cancel the input capacitive reactance at the grid. Although the input circuit is untuned, only 40 watts peak power is required to drive the amplifier in the frequency range 3.5-29.6 MHz. (The grid coils and switch aren't shown in the photo of the input compartment, since they were installed after the photo was taken.)

A Barker-Williamson coil assembly is used in the output circuit, with the 10-meter coil rearranged to shorten leads to the tube plate.

metering

Only two meters are used: a zero-center tuning meter and a plate milliammeter. Screen current is negative and does not require monitoring.

The zero-center meter permits quick adjustment for optimum drive and loading over any part of the band. It operates in the following manner:

The rf input and output circuits are sampled by capacitive dividers, then rectified and filtered. The dc outputs are connected to each side of a potentiometer, while the arm connects to ground through the zero-center meter. The diodes in this circuit are connected to provide a negative voltage from the input and a positive voltage from the output. The rf drive voltage then causes the meter to read to the left of zero, while the voltage from the output circuit opposes this action.

The 5-pF capacitor in the output sampling circuit is an adjustable wire probe coupled to the tank coil. The dc voltage at the 10-k zero-adjust pot can be adjusted by varying the position of this probe.

The voltage at the midpoint of the capacitive divider on the input circuit is about 3 volts peak (20:1 divider and -60V bias). The voltage at the midpoint of the output capacitive divider should be about the same.

In preliminary tuning, the usual procedure is followed using a two-tone signal and oscilloscope. Adjustments are made on drive and loading until the scope indicates minimum waveform distortion at the 2-kW PEP level. At this point the potentiometer is adjusted for a zero reading. Thereafter, when changing frequency, it's only necessary to resonate the plate circuit and adjust the drive and loading capacitor for a zero meter reading again.

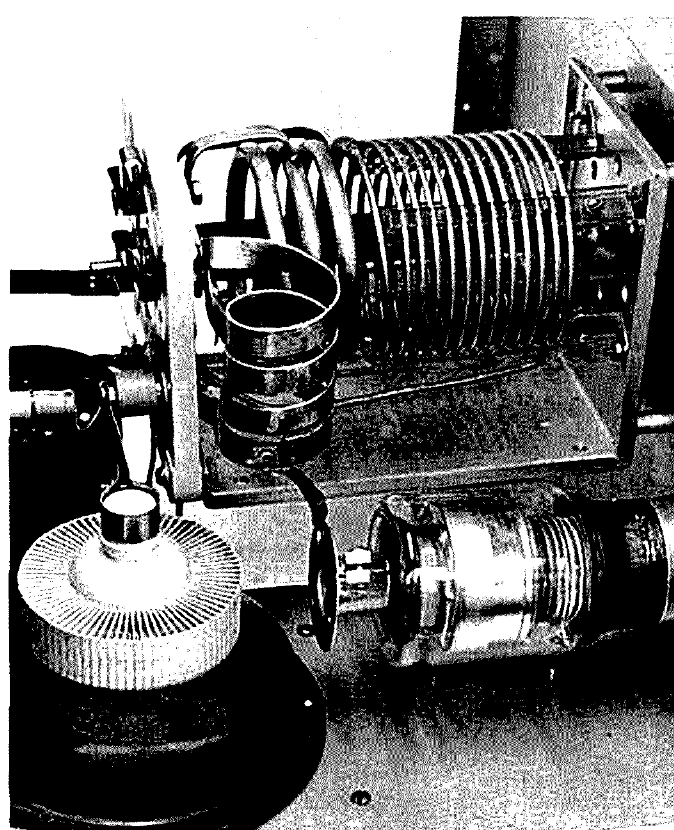
power supply

The power supply is conventional and uses silicon diodes in series in each leg of a bridge rectifier to deliver 2900-3200 Vdc, depending on line voltage. One-half of the secondary voltage at the transformer center tap is dropped by a resistor and connects to two series VR125 tubes, then to ground. This supply provides regulated 250 volts dc for the screen.

A single-section LC filter is used in the plate supply, and an electrolytic capacitor provides additional screen-voltage filtering.

A bias supply, using a small 115-volt

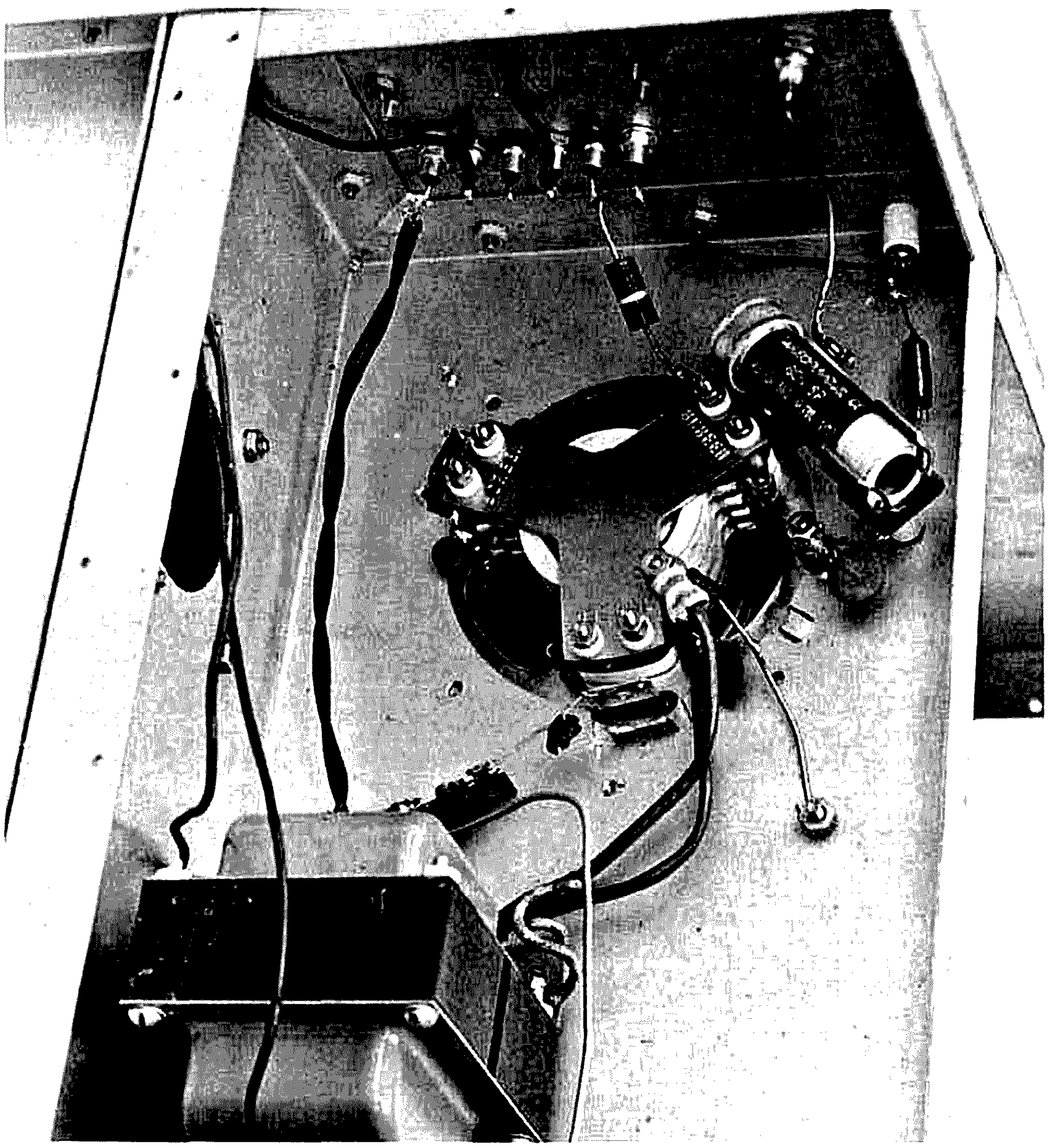
Tank circuit of the amplifier. The B&W 10-meter coil has been rearranged to provide short rf leads to the tube plate.



isolation transformer with a silicon bridge rectifier and RC filter, uses two parallel-connected potentiometers as voltage dividers. These pots supply bias voltages for "standby" and "operate" modes.

The auxiliary relay changes bias during standby periods to provide near cutoff plate current (50 mA).

With such light loading of the plate supply during standby, the plate voltage



Underchassis view. Grid circuit coils and line fuse were added after the photo was taken. Note arrangement of filament transformer leads and feedthrough capacitors for input voltages.

bias control

For good linearity, a plate resting current of 250 mA is required. During standby and talk periods this amounts to 800 watts plate dissipation, which runs up the power bill as well as the temperature of the shack. To remedy this situation, an auxiliary relay is connected to operate with the antenna transfer relay.

can soar to about 4000 volts, which imposes a strain on the filter capacitors. This problem was resolved by tuning the filter choke with a capacitor to approximately the supply ripple frequency of 120 Hz. With this circuit, a variation of only 200 volts exists between 50 mA and 250 mA loading. During "operate" mode, plate current varies between 250 and 350

mA. Overcurrent protection is provided by a relay in the negative side of the plate supply. A rheostat across the relay coil is used to adjust trip current. Circuit breakers are provided for the bias supply, ½-inch angles that form the box enclosure. The top cover of the amplifier is perforated aluminum to permit venting of blower air. The rear panel contains the 6-terminal board, coax connectors, and a

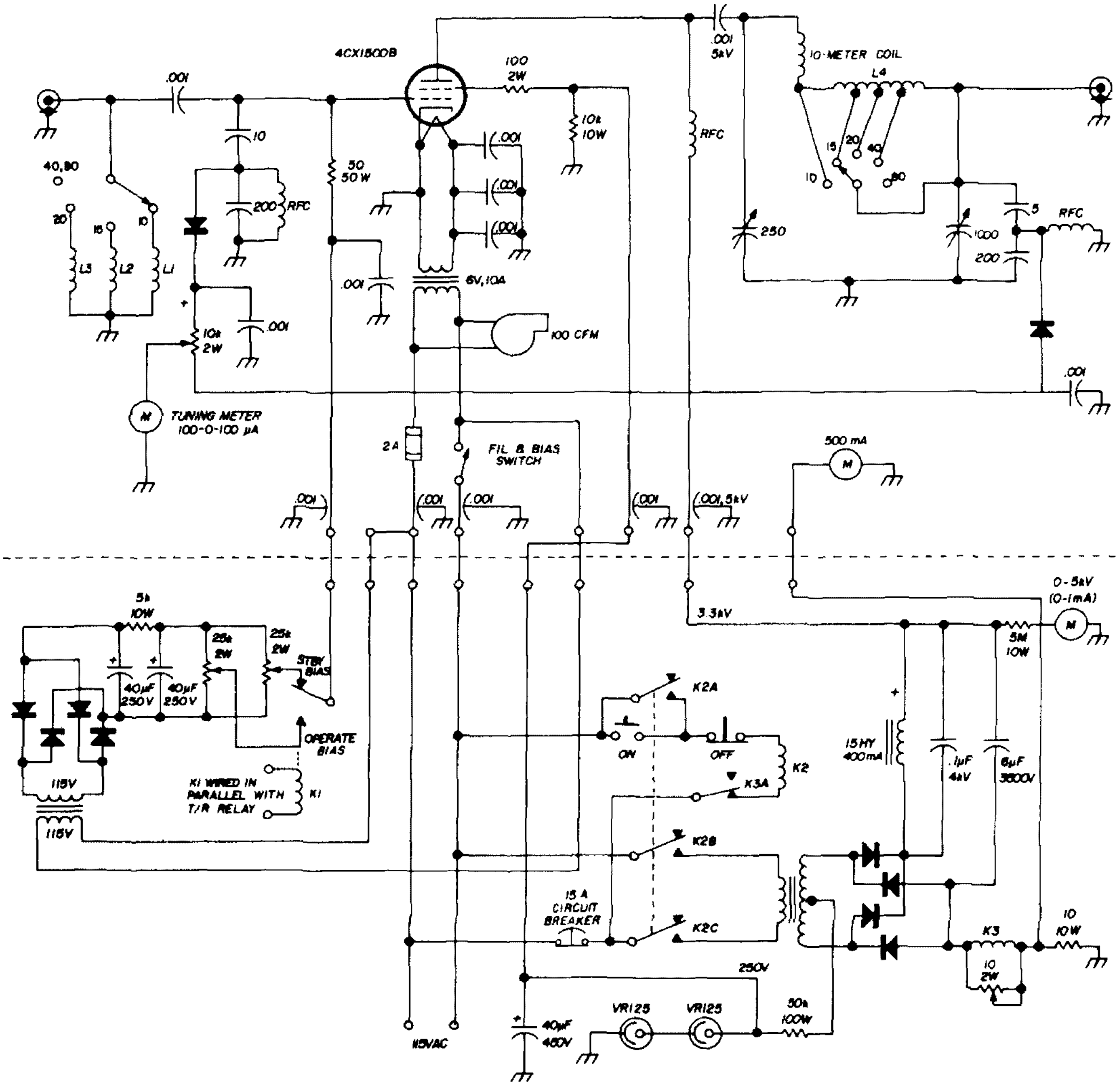


fig. 1. Schematic of the high-power 5-band amplifier. Grid coils are self-supporting, ½-inch dia., wound with no. 16 tinned copper wire. Ten meters, 5 turns 1¼-in. long; 15 meters, 7 turns 7/8-in. long; 20 meters, 13 turns 1-in. long.

filament transformer, and blower.

construction

Amplifier layout is designed for short rf leads. No neutralization or parasitic suppression is required. Aluminum plates are used, which are screw-fastened to panel fuse for the filament transformer. To minimize TVI, all electrical connections to the cabinet are made through feedthrough capacitors.

A tube chimney is used with the 4CX1500B socket to confine and direct blower air through the plate cooling fins.

tube characteristics

The maximum temperature rating for the 4CX1500B anode is 250° C. As an added precaution, it might be worthwhile to consider some sort of monitoring device for the tube. An article describing temperature sensors for high-power amplifiers, using scrs and thermistors, appears in an earlier issue of *ham radio.*[1] These circuits are extremely simple to build and are a good investment for the amateur interested in protecting his vacuum tubes.

heater voltage

Published ratings for the 4CX1500B heater is 6.0 volts. This means that heater voltage must be maintained within ±5 per cent as measured at the *tube socket terminals.* The heater voltage must be applied for at least 3 minutes before other operating voltages are applied.

screen grid

It is not uncommon for tetrodes to exhibit negative screen current. This is a prominent characteristic of the 4CX1500B and is normal for the circuit described here. However, the maximum rated power dissipation for the 4CX1500B screen is 12 watts, which should not be exceeded. Peak screen voltage times indicated dc screen current equals the approximate screen input power, except when screen current is near zero or negative.[2]

control grid

The control-grid dissipation rating of the 4CX1500B is 1 watt. Design features of this tube provide low intermodulation distortion, even when the grid is driven into the positive region. The tube can therefore be operated in Class AB_2, which is recommended in the manufacturer's literature.[2]

references

1. John J. Schultz, W2EEY, "Temperature Alarms for High-Power Amplifiers," *ham radio,* July, 1970, p. 48.
2. Data sheet, "4CX1500B Radial Beam Power Tetrode," Eimac Division of Varian, San Carlos, California.

ham radio

some observations regarding transmitter power levels

The difference between 100 watts and 1000 is not as great as you might think — on-the-air communications suffer little from lower power levels

Courtney Hall, WA5SNZ, 7716 LaVerdura Drive, Dallas, Texas 75240

The purpose of this article is to explain the ability of a 10-watt transmitter to span a continent with almost the same ease as a 100-watt rig. Experience has shown this to be true, and investigations have indicated that such phenomena should not be regarded as unusual.

In all of this discussion it is assumed that when *transmitter power* is changed, all else remains the same (frequency, antenna, vswr, receiver sensitivity, etc.). Also, power levels referred to can be input power or radiated power so long as it is the same in all cases; relative change and its observed effect are the main considerations. Transmitter efficiency and antenna gain and efficiency are assumed fixed.

on-the-air observations

Two years of operating experience with 50-watts cw seemed to indicate that I got out about as well as the fellows using 75- or 100-watt rigs. Although there was 50 watts difference between my rig and the 100 watters, this only amounted to 3 dB which seemed small in relation to a typical 140-dB path loss. The personal opinion grew that atmospheric conditions had a lot more to do with "getting out" than a few dB of transmitter power.

To gain more insight, I designed and built a small 12-watt, 40-meter cw transmitter which would deliver 8 watts output into a 50-ohm resistive load. I have used it for about six months on the air with the same antenna as was used for the 50-watt rig. This antenna is nothing special; just a 40-meter dipole about 15 feet above the ground, fed with RG-58/U. No balun is used.

It is difficult, if not impossible, to tell from the calls, QTHs, and signal reports in my log when I changed from 50 to 12 watts. I seem to get out nearly as well with only 12 watts. Since 12 watts is about 6.2 dB below 50 watts, I should average one S-unit lower in signal strength reports, but even this is not evident. I have received 559 to 599 reports from coast-to-coast, from Canada to Mexico City on 40-meters cw with both rigs. Furthermore, I have not noticed any greater difficulty in making contacts with the 12-watt rig.

laboratory observations

Still, I had no feeling for how a signal change of 3 dB, 10 dB or 20 dB sounded. To find out, I assembled the equipment shown in fig. 1.

The 1-kHz audio oscillator had a sine wave output, and its level control allowed the output to be adjusted from zero to approximately 10 volts rms. A key was used to simulate actual code reception, rather than a steady tone. The step attenuator was a commercial type which could be adjusted from 0 to 120 dB in one-dB steps. To achieve a close match

for the output impedance of the attenuator, an 820-ohm resistor was connected in parallel with the 2000-ohm phones.

To begin the experiment, the attenuator was set for zero-dB attenuation and the oscillator level control was adjusted for a comfortable volume level in the phones, such as would be used for code practice. Then, 3-dB attenuation (half power) was switched in on the attenuator; this caused a slight decrease in volume, but it would be extremely difficult to distinguish between the two levels without being able to switch between the two. In other words, if you listened to one level on one pair of phones, then took those phones off and put on another pair of phones to listen to the other level, they would sound almost the same.

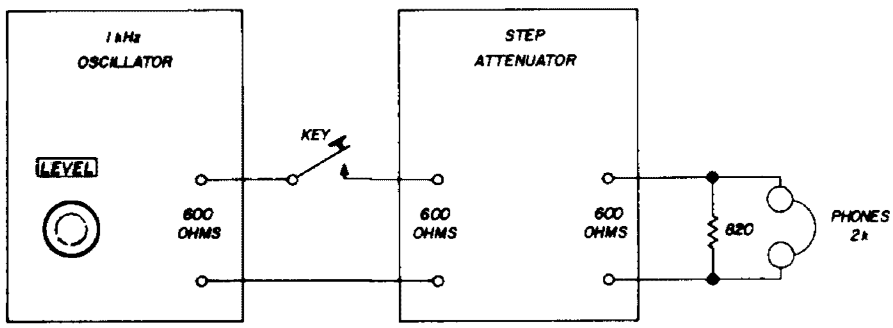


fig. 1. Bench test circuit for determining the effect of different power levels.

Next, the change in volume corresponding to 10-dB attenuation (one tenth power) was observed. The signal was noticeably weaker, *but not so weak that you would take the trouble to increase gain if you were practicing code.* It was still a good signal, just a little below optimum. This observation explains, to me at least, why a 10-watt transmitter will sound fairly good to a receiving station if a 100-watt transmitter sounded very good, all else being equal.

Attenuation of 20 dB (reduction of power by a factor of 100) made the signal pretty weak, but it was still easily readable; 30-dB reduction was extremely weak and difficult to read, and 40 dB was barely perceptable. When the signal was increased to 10 dB above the original zero dB, or comfortable level, the volume was uncomfortably loud and almost painful.

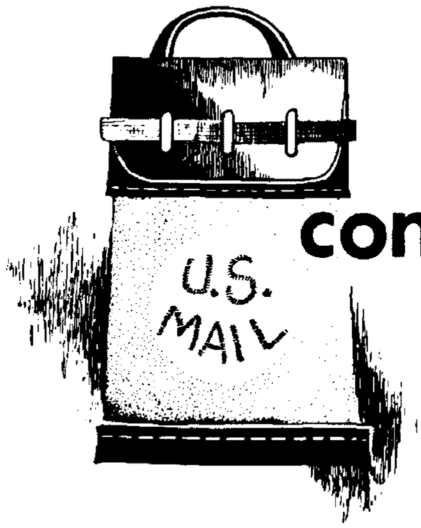
I believe the above observations show that the novice limit of 75 watts is far in excess of the number of watts required for casual cw work, and that there is no need to feel ashamed of your 100-watt rig if a 10 watter gets as good a report as you do; 10-dB difference in power may represent a lot of watts and dollars, but it doesn't impress the ear that much.

To put all of the above in its proper perspective, the following should be mentioned. It's not the easiest thing in the world to describe in words how loud a sound is; I have tried to describe the laboratory observations as I heard them, but the dynamic range of my ears may be average, above average or below average. Much depends on the "comfortable" level you start with at zero dB. The laboratory observations simulated ideal conditions; there was no receiver noise, no interference and no fading.

I have no argument with the man who says the more power you have, the better chance you have to get through. He is absolutely right. My point is that you can do more with just a few watts than perhaps many hams realize.

It is interesting to express the ratio of Novice power limit to General power limit in dB: 11.25 dB. Another point to remember is that receivers with cw agc tend to make the 10- and 100-watt rigs sound the same, the difference appearing on the S-meter. Finally, the laboratory observations should give an indication of what to expect from filters; at least 30 dB of attenuation should be provided for unwanted frequencies which have a signal strength equal to the desired signal.

ham radio



comments

motorola fets

Dear HR:

I enjoyed reading the September issue of HR – especially the piece on the fm receiver. Great!

I noticed that the authors referred to an MPF105 and MPF107 as being better than an MPF102 in the front end of the unit. Tain't so! The MPF105 is rated to a much lower frequency than 2 meters. The MPF107, however, is good to 3/4 meters. Perhaps he meant to say MPF106 and MPF107 there. At any rate, the MPF designation has been obsolete on all but the MPF102 series for almost a year now. This could give your readers a bit of a problem. Here are the new Motorola numbers for the devices:

MPF102 . . . no change
MPF103 . . . 2N5457
MPF104 . . . 2N5458
MPF105 . . . 2N5459
MPF106 and 7 now split into three types: 2N5484 through 2N5486

Also, there is a new series of MPF102 devices – 2N5668, 69, and 70. These are simply more tightly controlled MPF102s.

Doug DeMaw, W1CER
ARRL Technical Editor

injection lasers

Dear HR:

The excellent article on injection laser experiments by Ralph W. Campbell (November, page 28) was of particular interest to me. I am convinced that lasers have a definite role in amateur communications, and it is encouraging to see experimenters such as Mr. Campbell pioneering the way.

Regarding communications with modulated light emitting diodes, I would like to point out that Henry E. Roberts and I have covered the subject in some detail in a pair of articles in the November issue of *Popular Electronics.* One article is a tutorial, and the other is a construction project for a 1,000-foot range LED voice communicator, The latter is being marketed in kit form by MITS, Inc. Price of transmitter and receiver, both of which are quite compact and battery powered, is $32.00 postpaid. For more information, write the company at 4809 Palo Duro, NE, Albuquerque, New Mexico 87110.

While the light emitting diode is actually superior for many voice communications experiments at the present time, the laser does have some special benefits, particularly peak power. For those experimenters desiring to work with semiconductor lasers, I would discourage the purchase of the RCA TA-2628. The TA-7606 uses a new technology and is far superior. For example, the latter laser has a threshold *for lasing of about* 4 amps – considerably less than the

TA-2628. Also, lifetime will be longer with the newer laser. There are numerous other considerations and the experimenter will be wise to consult some of RCA's excellent data pamphlets, such as "GaAs Lasers and Emitters" OPT-100, before actually selecting a laser. As these devices can be "zapped" quite easily, particular care to circuitry requirements should always be observed.

Forrest M. Mims
Albuquerque, New Mexico

fm receiver frequency control

Dear HR:

Several weeks ago I acquired a surplus Motorola base station. The receiver was easily aligned to 146.94 MHz, the national calling frequency. Sensitivity was better than 0.1 easy microvolt to crack the squelch, and 0.3 for 20-dB quieting. (The signal generater was an HP-608F.)

The set uses a 12.455-MHz second oscillator and a 455-kHz Permakay filter. This filter (TU-540S) is narrow – about 10 kHZ. Even after tuning the first oscillator for nominal input-frequency, more than half the signals, sufficient to open the squelch, are unintelligible, This is for three reasons: Low levels, as shown by monitoring an i-f grid-return; Signal deviation too wide for filter in use; And signal frequency off center of response. A considerable proportion of unintelligible receptions was due to the last, which is the subject of this note. The problem is well known on the 470-MHz commercial band, and is handled in part by afc to the first crystal oscillator. In the 160-MHz band the practice seems to depend on oven-mounted first-oscillator crystals, and frequent touchup of fleet frequencies by a service shop.

Compared to afc, receiver standby with "un-rubbered" quartz has some advantages. The center response can be precisely set with a counter, and tends to stay put. Application of afc to crystals is not so easy at 160 as at 470 MHz. In any case, it degrades stability.

Because of spectrum congestion, trends are to split channel allocations. In two-way fm, 60 to 30 to 15 kHz. In two-way aeronautical service, a-m, 100 to 50, and likely, 25 kHz. How to deal with narrow-band stability problems seems to be of both commercial and amateur interest.

Why not use the crystals for standby, and automatically transferring to afc when receiving a carrier sufficient to open the squelch? In fm receivers the necessary discriminator and amplified squelch already exist. For a-m receivers these elements have to be added. To be added in either case are switching circuitry controlled by carrier/noise-operated voltage, and a vco. To impose lesser demands on inherent vco stability, the second oscillator looks like the place to enter. As mentioned, this is at 12-MHz in a typical case. Transfer time-constants must meet various criteria, one being that reception must be continuous to keep the squelch open. Ideally, the vco would start, no-break, from synchronism with the crystal, and proceed within a few multiseconds to discriminator zero.

Has this been done already? Where can I get details?

Paul D. Rockwell, W3AFM
Chevy Chase, Maryland 20015

digital counter

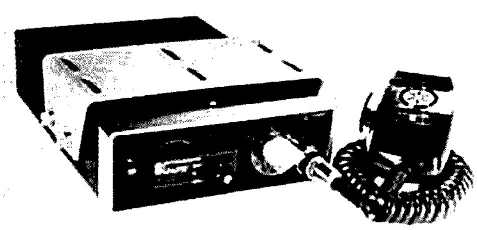
Dear HR:

I wish to thank you for a very fine magazine, in particular the articles by Bert Kelley, K4EEU. I built the digital frequency counter described in the December, 1968 issue, and used it in the September, 1969 and February, 1970 ARRL Frequency Measuring Tests. It enabled me to be among the top group who had accuracies better than 0.4 parts per million on both tests. In fact, my readings coincided exactly with the umpire on the February, 1970 run.

Joseph Czerniak, W8NWU
Muskegon, Michigan

new products

fm transceiver



Standard Communications Corporation, the largest manufacturer of vhf marine equipment in the world, has developed a professional quality vhf-fm transceiver specifically for amateur use. Standard has applied the latest solid-state technology to build a unit that is rugged and compact, and makes mobile installation practical in nearly any vehicle or aircraft. In addition, it's fully portable with the addition of a battery pack.

The Standard model SR-C826M features 12 crystal-controlled channel capability (4 channels factory installed), 10 watts rf output power, low power consumption, mosfet rf amplifiers and mixer, noise-operated squelch, high-low power switch, illuminated dial, self-contained speaker, and metering of battery voltage, relative received signal strength and power output. Current consumption of the two-meter transceiver is 150 mA on standby and 2.4 amp on transmit. Supply voltage is 11 to 16 Vdc, 13.8 Vdc nominal. Frequency stability is 0.001% from -10 to $+60\,^{\circ}$C.

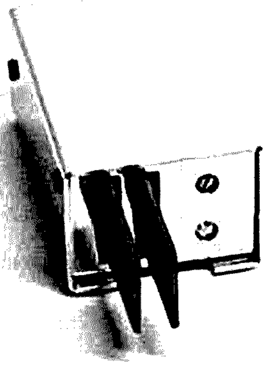
Transmitter power output is 0.8 or 10 watts, depending on setting of hi/lo power switch. Output impedance is 50 ohms nominal. Frequency deviation, internally adjustable to ± 10 kHZ minimum is factory set to ± 7 kHz. Spurious and harmonic responses are attenuated 50 dB below carrier power level. Audio rolloff above 3 kHz at 16 dB per octave.

Receiver sensitivity is 0.4 μV or less for 20-dB quieting. Squelch threshold is 0.2 μV or less; maximum (tight) squelch between 20=dB quieting sensitivity and 20 dB quieting plus 10=dB. Deviation acceptance up to ± 15 kHz. Spurious and image attenuation 65=dB below the desired signal. Adjacent channel selectivity (30-kHz channels) is 60 dB attenuation of adjacent channel. Audio output, 5 watts minimum (with external speaker); audio distortion, 10% maximum at 3 watts output. Priced at $399.95 complete with microphone, built-in speaker and external 2.5 amp alternator whine filter. For more information, write to Standard Communications Corporation, 639 Marine Avenue, Wilmington, California 90744, or use *check-off* on page 94.

ferrite cores

Indiana General CF102-Q3 toroid cores are now available from HAL Devices for $1.25 each. (This core was used in the rf bridge featured in the December, 1970 issue of *ham radio.*) Other ferrite toroids available from HAL are the Indiana General CF102-06, CF102-Q1 and CF101-Q2; these cores are $.50 each. In addition to ferrite cores, HAL Devices carries an interesting line of hard-to-get products for the experimenter, including integrated circuits, hot-carrier diodes, integrated-circuit sockets and Mainline RTTY equipment kits. For more information, use *check-off* on page 94, or write to HAL Devices, Box 365H, Urbana, Illinois 61801.

deluxe permeflex key



James Research has announced a new wider and heavier version of the Permeflex key. The completely enclosed mechanism has independent fiberglass paddles which flex to make contact and have adjustable gap and tension. Additional feet allow the key to be used on its side as a straight hand key. The 8-ampere gold-diffused contacts may be used to key a transmitter directly, or to key the low-level input of an integrated-circuit keyer.

Included with the new deluxe key is an internal bracket and blank printed-circuit board to permit construction of a built-in electronic keyer or monitor. The pre-punched 1/4" holes on the front panel will accommodate subminiature switches and potentiometers. The key is guaranteed for 1 year. $24.95 postpaid in the U. S. A. For more information, write to James Research Company, Department HQ-K, 20 Willits Road, Glen Cove, New York 11542, or use *check-off* on page 94.

alpha-seventy linear amplifier

The Alpha Seventy is a new generation high-power linear amplifier that incorporates many major state-of-the-art advances in design. It easily handles the maximum legal amateur power in any mode without any "on-time" limitations, yet is completely self-contained in a desk-top cabinet. Among the many features of this unique new design is *quiet operation* through the use of vapor-phase

cooling (no high-velocity fans) and quiet vacuum-sealed transmit-receive relays.

The Alpha Seventy features cool operation because the vapor-cooled Eimac 3CV1500A7 tube reaches only 100° C under full load, not 200° C like most air-cooled tubes. Heat is flushed out the rear of the cabinet, away from the operator and other equipment.

Other features of the Alpha Seventy include a built-in mosfet electronic T/R switch that is preselected on receive by the extremely high unloaded Q of the plate tank circuit, cutoff bias that is electronically controlled to prevent plate current flow in the absence of rf drive, dc coupled alc output that simplifies tuneup and provision for later installation of an accessory (to be announced) to provide remote-control or automatic band change, tuning and loading.

The Alpha Seventy covers all amateur bands from 80 through 10 meters; a range extension kit for 160 meters is to be announced. Drive requirements are 100 watts PEP nominal for 2.5 kW PEP plate input; 50 watts nominal for 1 kW carrier input. Third-order products are more than 30 dB below each of two equal test tones at 2.5 kW PEP input.

Priced at $1595. For more information, write to Ehrhorn Technological Operations, Inc., Post Office Box 1297, Highway 50 East, Brooksville, Florida 33512 or use *check-off* on page 94.

160-meter antenna tuner

The Top Band Systems model 48MV 160-meter antenna tuner is designed to resonate practically any 40- or 80-meter dipole or inverted vee on 160 meters. The unit covers the entire 160-meter band, and can be used with coaxial cable or open-wire feedline. The tuner will handle 140 watts cw or a-m, or 250 watts PEP ssb. The 48MV Matchverter can also be used to resonate long-wire antennas. $39.95 from Top-Band Systems, 5349 Abbeyfield Street, Long Beach, California 90815, or use *check-off* on page 94.

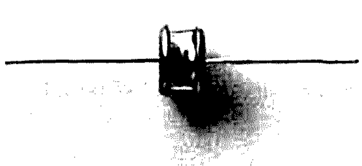
semiconductor cross-reference catalog

Motorola's HEP sales department has announced a new cross-reference and replacement guide, HEP HMA-07, available free through HEP distributors nationwide. This guide cross references more than 25,000 devices to HEP replacements including 1N, 2N, 3N, JEDEC, Japanese, *Dutch* and other foreign numbers in addition to thousands of manufacturers' regular and special "house" numbers. There is special emphasis on replacement coverage of the device numbers found in consumer products equipment, particularly Japanese merchandise, and several thousand industrial MRO market device types.

This new semiconductor cross-reference guide and catalog includes Motorola's full-line product catalog, which gives the min/max ratings and electrical characteristics for 285 HEP devices, as well as cross-reference information. The Motorola HEP devices are listed by type number with a packaging index, device dimension drawings and selection guide information.

HEP is Motorola's sales program for making semiconductor devices readily available to the experimenter through a nationwide network of authorized suppliers.

surge voltage protectors



Two new miniature gas-filled surge-voltage protectors to safeguard inputs of electronic instruments as well as semiconductors and other components against

lightning, power surges, and other transients, have been developed by Siemens.

The new units, type B2-H10 and B2-H25, have two features of particular advantage to electronic component users and designers: they are only 0.28- and 0.44-inch long, respectively, and provide protection for equipment with peak operating voltages up to 850 and 2000 volts, respectively.

Ultrafast in response, the SVP type B2-H10 has a dc striking voltage of 1000 volts (±15%); for the SVP type B2-H25, the dc striking voltage is 2.5kv. Insulation resistance is greater than 10,000 megohms, 5000 amps rated discharge capability for a 0.3-microsecond wave, and capacitance of less than 2 pF.

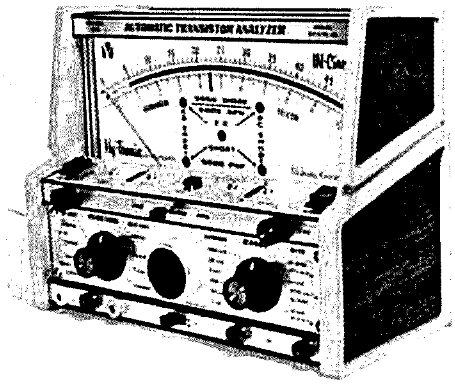
The two new SVPs are part of a complete line of surge voltage protectors now being marketed by Siemens. Additional information is available from Siemens Corporation, 186 Wood Avenue South, Iselin, New Jersey 08830, or use *check-off* on page 94.

low-noise uhf transistors

The Nippon Electronic Company has announced a new family of low-noise silicon transistors designed for high-gain applications in the vhf/uhf region below 1000 MHz. These new devices are suitable for broadband amplifiers, rf and i-f preamplifiers, oscillators and digital circuits requiring ultra-fast switching.

The gain-bandwidth product, f_T, for these transistors is typically 3000 MHz; power gain at 500 MHz is typically 25 dB. The 2N5652, with a 1.8 dB typical noise figure at (2.5 dB maximum) 500 MHz, is $11.40; the 2N5651, NF 1.3 dB typical (2.0 dB maximum) at 500 MHz is $25.80; the 2N2650, NF 1.0 dB typical (1.5 dB maximum) at 500 MHz is $53.50. Prices are for small-quantity orders. Units may be purchased from California Eastern Laboratories, Inc., 1540 Gilbreth Road, Burlingame, California 94010. For more information use check-off on page 94.

automatic transistor analyzer



Vanguard Electronic Tools has introduced a new automatic transistor analyzer, the model 900, which measures current gain (beta) directly, and measures leakage to the nearest nanoamp. This analyzer, designed for use in or out of the circuit, automatically differentiates between pnp and npn transistors, indicates silicon or germanium, and operates on 115/230 Vac.

Current-gain measurements are zero to 500 (at 20 μA and 200 μA I_B) and zero to 50 (at 200 μA and 2 mA I_B). Leakage can be measured at both I_{CEO} and I_{CBO}. The leakage scale in the 1-μA position has 10 nanoamps per minor division. Collector current is indicated from 10 μA through 100 mA, and the emitter-base voltage drop can be measured to the nearest millivolt.

The in-circuit transistor test is fixed at the 270-ohm shunt level, which means that as long as the test transistor has more than 270 ohms across the junctions, the model 900 will see only the test transistor, and none of the peripheral circuitry.

For qualitative go-no-go testing of transistors and diodes, an audio readout can be activated that will indicate all good transistors and diodes. This is especially helpful when the operator finds it impossible to watch the meter. The price of the automatic transistor analyzer is $287 (less model 103 electronic probe, which is $15). For more information, write to Vanguard Electronic Tools, Inc., Hy-Tronix Instruments Division, Post Office Box 667, Newton, Kansas 67114, or use *check-off* on page 94.

short circuits

fm receiver

There are several corrections to the fm receiver article that appeared in the September, 1970 issue. Inductors L201, L203, L204 and L205 should use number-24 wire, *not* number 18. Inductors L101 and L102 are Nytronics part number SWD5600. The 22k resistor across T106 can be changed to 100k to facilitate oscillation if T106 has low Q.

Transistor Q204 may not provide sufficient injection to the mixer due to gain differences, manufacturing tolerances, etc. The injection level can be increased by replacing Q204 with a Fairchild SE3005. Be sure to retune T202 and L206 after making the replacement; it may be necessary to add a few more pF across T202 to obtain resonance with the SE3005.

If it is desirable to have a "squelch tail" on your receiver, connect a 10-μF capacitor from the base of Q107 to ground.

QRP indicating wavemeter

In fig. 1 on page 27 of the December 1970 issue, the 1-mA meter will be zapped if the switch is closed. The battery should be returned to ground through the switch to prevent damage to the instrument.

frequency scaler

On page 28 of the August, 1970 issue the polarity of the electrolytic capacitors are installed backwards; the rectifier bridge is shown correctly. One reader has noted that the photograph shows the electrolytic backwards; this, according to the author, is due to an electrolytic casing being incorrectly installed by the capacitor manufacturer.

inexpensive swr indicator

There are a number of false statements in this article that got by our staff, including misinformation about transmission lines, antenna couplers and impedance matching. If you haven't read this article yet, ignore it; if you have, forget it.

Rf, in coax, flows on the *outside* of the inner conductor, and on the *inside* of the outer braid. There should be no rf on the outside of the outer shield (except at microwave frequencies where the braid leaks). If there is, it is usually caused by pickup from antenna radiation or the fact that an unbalanced line is used to feed a balanced antenna, and rf is coupled directly from the antenna to the outside of the outer braid. This rf sets up parallel standing waves and this is what is detected by the device described in the article. Our staff makes every effort to weed out technical inaccuracies, but we slipped badly on this one, and have re-evaluated our procedures to insure that we don't publish misleading and false information in the future.

BC-1206 conversion

The 500-ohm resistor connected across the 250-pF capacitor in fig. 1 on page 32 of the October, 1970 issue should be 500 k.

fm frequency meter

The 60-kHz output in fig. 2, page 42 of the January, 1971 issue should be connected to the emitter of the 2N2219, *not* the collector.

ST-6 Mainline RTTY demodulator

In fig. 6, page 18 of January, 1971 issue, there should be a lead from the input connector to the large dot between R1 and the 620-ohm resistor.

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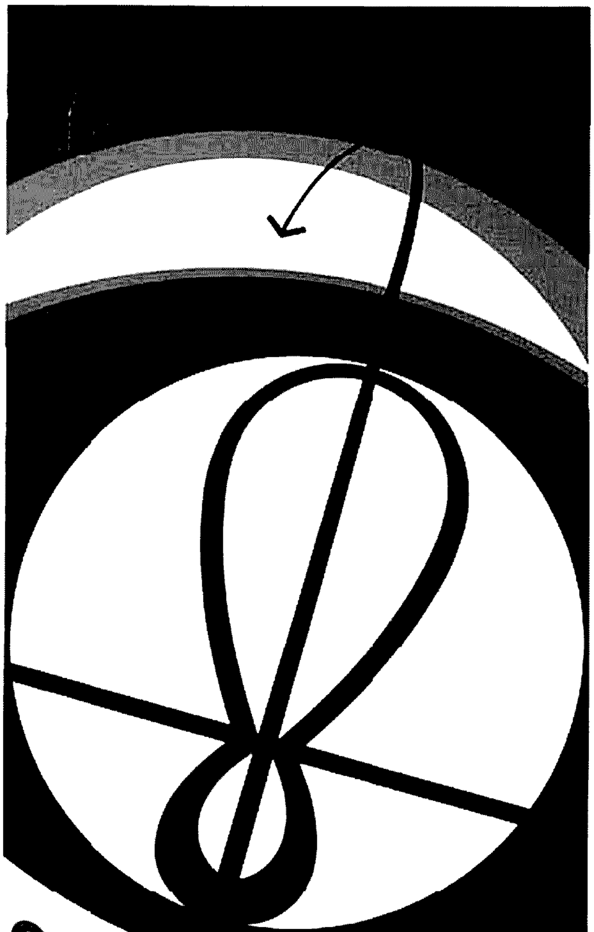
MAY 1971



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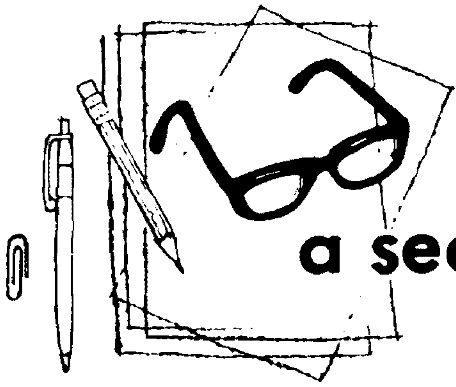


a second look

by jim fisk

With the coming of late spring it is time to take a close look at your antenna system to make sure it's operating up to par. Check *all* the *electrical* connections for oxidation, inspect the feedline for damage, and if the swr is not as low as it used to be, now's a good time to tune up the system.

If you are considering a new antenna for your station you may find useful one of the designs in this, our annual antenna issue. Although there have been few important breakthroughs in amateur antenna design since the cubical quad, a great many specialized antennas have been developed for military and space communications.

One recent example is the log-periodic, first described as a microwave antenna, later as a high-frequency wire beam, and finally, for television broadcast reception. The log-periodic antenna has never become especially popular with amateurs, probably because we operate on segmented bands, and the log-periodic is a broadband device. The same might be said for other recent developments such as the backfire antenna and the phased array.

Over 100 years have elapsed since Maxwell formulated the celebrated mathematical equations that continue to provide the basis for classical radio theory. This theory was first verified by Hertz in 1887 with a center-driven wire antenna, 24 inches long, terminated at each end with a metal plate about 16 inches square. The fundamental frequency of this antenna, 53.5 MHz*, was excited by a spark gap.

Marconi's first successful wireless experiments used basically the same antenna: Two large copper plates excited by a spark gap. To increase the distance of *his transmissions* Marconi put up larger and larger quantities of wire. Early amateur operators did the same – the DX performance of a station was often directly proportional to the size of the station's antenna.

The first advances in gain antennas were the result of experiments by commercial radiotelephone companies which wanted to improve the reliability of their overseas service. First came the long-wire, the vee and the rhombic; then the Lazy-H, Sterba curtain and Bruce array, followed closely by the W8JK flat-top and the parasitic beam. Each new design was eagerly tried out by amateurs trying to improve the performance of their stations.

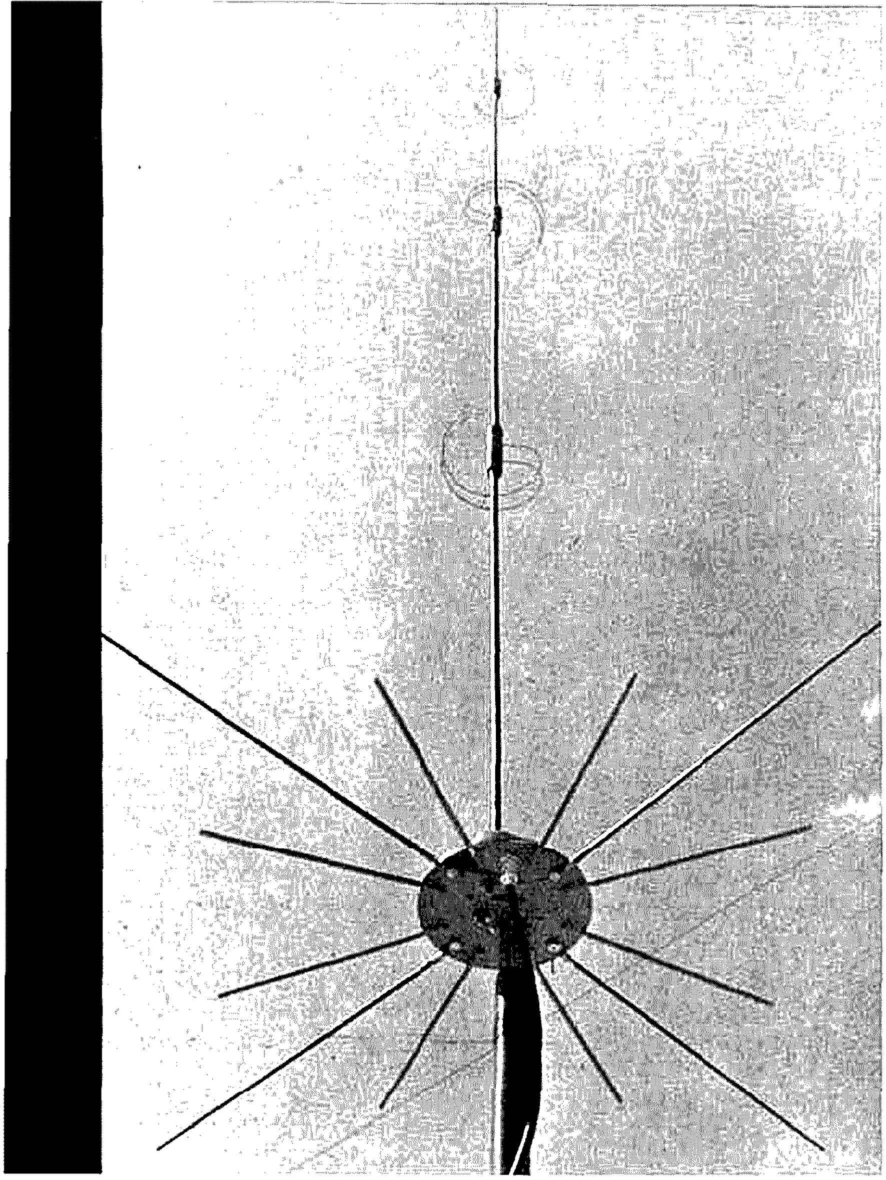
However, today it is different; the antennas are not that much different from those used a decade ago. Feed systems have been improved, impedance matching is better, and the antenna will stand up under more severe weather conditions, but the basic antenna is little changed. I wonder how much more we can improve the ubiquitous beam antenna?

expanded phone bands

The actual FCC docket for expanded phone bands is essentially the same as that discussed on this page last month. The text of Docket 19162 is printed in its entirety in the April issue of *QST*. If you don't have access to a copy of the text we will be glad to send you a copy if you will enclose a self-addressed, stamped envelope with your request.

Jim Fisk, W1DTY
editor

* R. King, "The Linear Antenna – Eighty Years of Progress," *Proceedings of the IEEE*, January, 1967, page 2.



four-element colinear array for two meters

Extended elements with phasing stubs give 14 dB gain over a ground-plane antenna

Bob Dahlquist, WB6KGF

Here is an antenna with as much gain as a medium-sized yagi beam; yet, unlike the yagi, this antenna is omnidirectional. You can easily work stations from all points of the compass without waiting for a rotator to crank the antenna around to the desired azimuth — an advantage when working in a contest, a net, or with mobile stations.

the circuit

The antenna is known as a 4-element extended colinear array. It's an improvement over the classical colinear antenna described in the literature.[1] Instead of ½-wavelength elements, the extended colinear uses 5/8-wavelength elements

(except for the topmost element) to obtain gain over the conventional colinear array or the popular ¼-wave ground-plane antenna.

The topmost element is shorter than the other elements to preserve current balance in the phasing stubs. The sketches in **fig. 1** show the electrical circuits and the distribution of current and voltage.

have? I made a comparison test, starting with a simple ground-plane antenna. When the extended 4-element colinear was substituted, the radiated field strength increased by 14 dB. Some of this increase could be attributed to an increase in power input to the colinear, since the colinear was adjusted to near unity swr before it was installed.

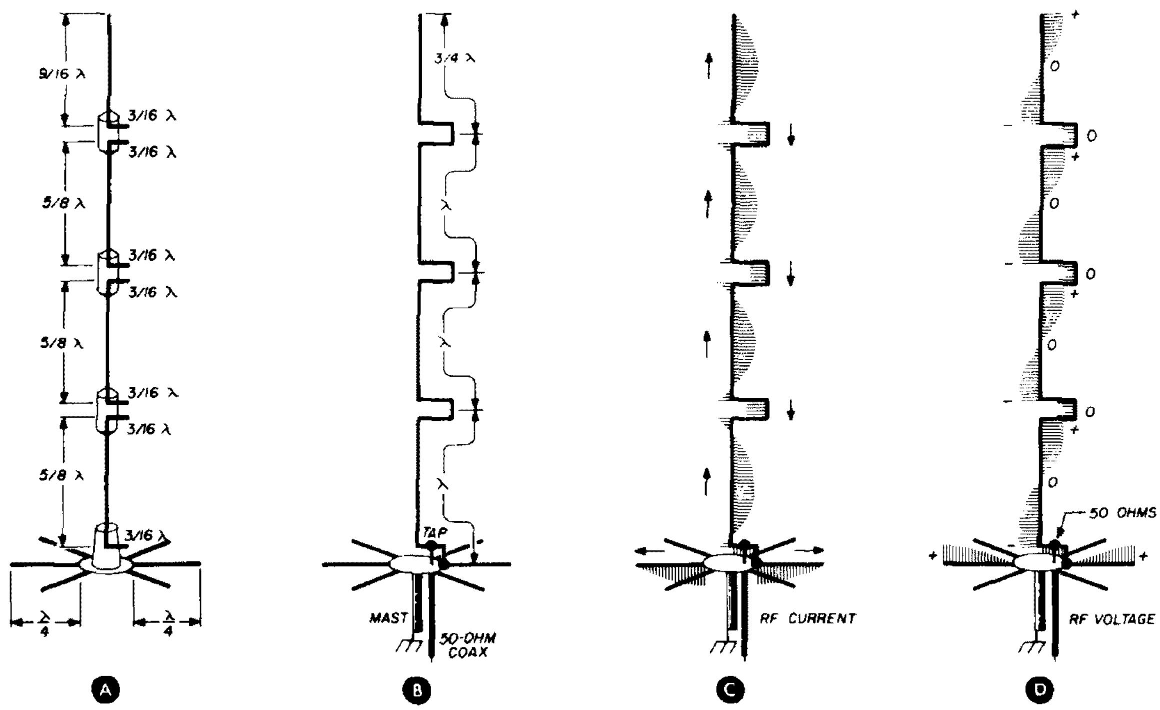


fig. 1. The extended colinear array. Sketch A shows electrical lengths of elements and tuning stubs; B gives the electrical circuit of the complete antenna. Current and voltage distribution are shown in C and D.

features

In addition to the added gain, the extended colinear has extra height compared with the ground plane. The extended colinear is self-supporting to 17 feet above its base on 2 meters. This added height puts the radio horizon several miles further away than that of a ground-plane antenna. The improvement will be more noticeable, of course, if you must roof-mount the antenna than if you can manage a 50-foot tower.

gain

Just how much gain does this antenna

theory

The extended colinear array is a bit longer than the classical 4-element colinear. The added gain results from the increased spacing between current loops, or maxima, in the radiating elements (**fig. 1**). Increasing element length beyond 5/8 wavelength actually decreases gain, because the out-of-phase currents in the element ends then become large enough to cancel some of the radiated field.

phasing

By mounting the four elements linearly and feeding them in phase, the fields

radiated by the individual elements will reinforce. The sum of the four fields (total field strength) will be maximum in a plane normal to the axis of the antenna (fig. 2). The pattern is like the doughnut-shaped field of a vertical dipole, but much flatter. In the vertical plane, the beam-width is only about 20 degrees, with maximum radiation directed at the horizon (fig. 3).

Proper phasing is accomplished with stubs. Placed between each element, a stub of the proper electrical length introduces just enough delay to keep the currents in the elements phased 360 degrees apart (for our purpose, this is the same as zero degrees, or exactly in phase).

Since the elements are longer than usual, the phasing stubs must be shorter by a corresponding amount. Instead of ¼ wavelength, the stubs are 3/16 wavelength from the shorted end to the open end, or 3/8 wavelength all the way around. The length of one element, plus the length of one side of each adjacent stub, equals one wavelength (360 electrical degrees) except for the top element, where the total is ¾ wavelength; the length of the top element is 9/16 wavelength (see fig. 1). Lengths of all elements, tuning stubs, and radials are given in fig. 4.

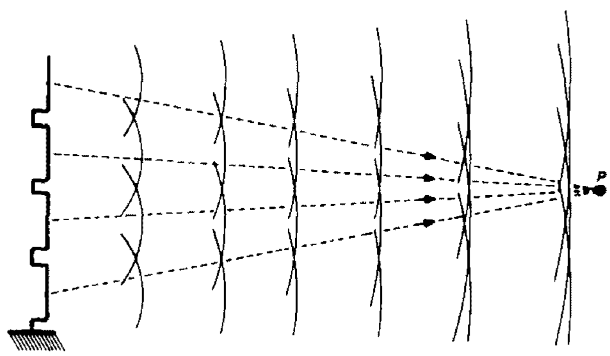


fig. 2. Relationship of waves radiated from each element as a function of range. Since the waves travel approximately equal distances to reach a distant point, P, all arrive in phase and add to produce increased field strength.

feeding and matching

The array is fed at the base through another stub, much like the phasing stubs, except that only half of this stub is constructed. (The other half is a reflection in the ground-plane disc.) Since this stub is unbalanced, coax can be connected directly without a balun. The center conductor is passed through the disc and soldered to the point on the stub

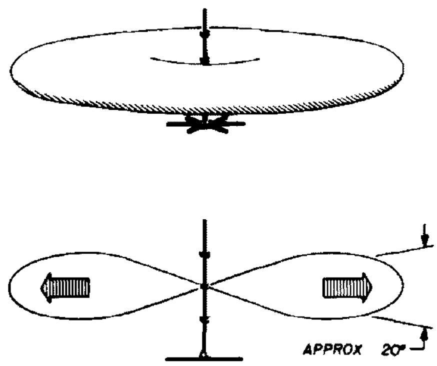


fig. 3. The "squashed doughnut" pattern of the extended colinear antenna. The angle shown is in the vertical plane.

where the impedance matches that of the coax, and the braid is grounded to the disc that secures the ground-plane radials. See fig. 5. By alternately adjusting the position where the coax is tapped, and the length of the stub, a match near 1:1 can be obtained. The radials are ¼ wavelength long. They decouple rf from the mast and the outside of the coax, thus making the array independent of actual ground. An overall view of the antenna is shown in fig. 6.

construction

This antenna is best assembled by casting the insulators in place. This procedure is not absolutely necessary, but it does produce strong, watertight joints. The casting process is not difficult or expensive; one method is described later in the article.

radiators

Each radiating element should be

strong enough to support both the weight and the wind load of the parts above it. I used tapered elements; it would also be possible to make each element from a different-size tubing. Aluminum alloy is a good choice for material because of its high strength-to-weight ratio, although it's difficult to solder.

If each element is made with a joint of some type, the antenna can be disassembled to make transportation easier. If this isn't done, the antenna should be assembled in a place where its 17-foot length won't become a problem. My tubing had threaded joints; however, lacking these, you can use progressively smaller sizes of tubing. By splitting the end of a tube with a hacksaw, slipping the smaller tube inside and applying a hose clamp or two, you'll have a joint that will allow adjustment as well as disassembly of each element.

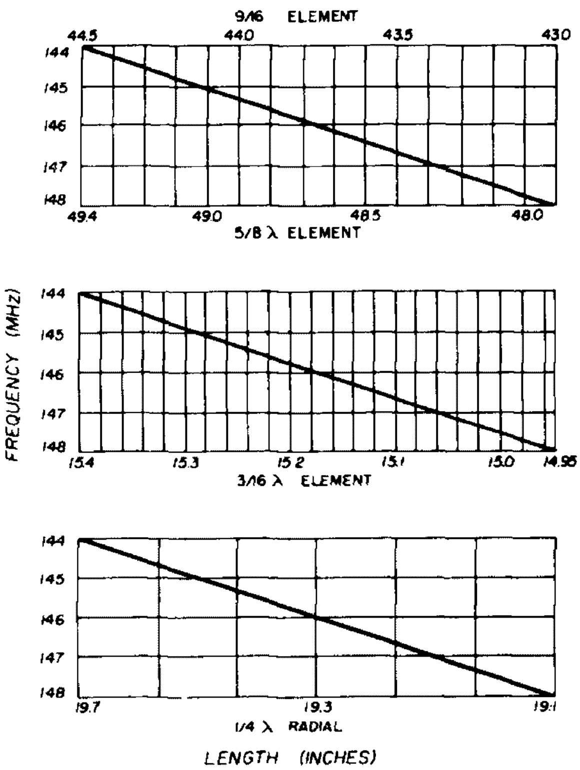


fig. 4. Lengths for elements, stubs, and radials in inches versus frequency. Data is based on 1/2 to 3/8-inch-diameter elements.

Each stub is made in two halves at first, to be joined later. Cut each half to length according to **fig. 4**, but allow a little extra for making the joint to the tubing. Add a little more than this for pruning; say 6 inches. It's best either to solder or weld the wire to the tubing. Make sure all joints are mechanically and electrically sound. Do *not* use acid-core solder.

casting the insulators

For casting, you'll need three basic items: a mold, resin, and a means to hold the parts steadily while the resin cures.

The mold can be made of plaster of Paris. You can make a mold for each insulator and do the casting all at once, or you can make one reusable mold and cast the insulators one at a time except, of course, the base insulator, which requires its own mold. I used the latter method; however, if I were to do it again I'd use the former method, which saves a lot of trouble and resin.

element insulators

Arrange each element with stub wires attached, and with the joints disassembled, so that each element is in its proper relative position with respect to the others. A cardboard mold, fashioned from a quart milk carton, is then used to cast a block of paraffin. When the paraffin has solidified, remove the cardboard mold and carve the paraffin into the shape of the insulator (see photo).

Using another piece of cardboard, cast a block of plaster of Paris around the wax insulator. This will form a mold for the plastic resin, which constitutes the insulator. When the plaster has set, the wax can be melted. You'll need at least one hole in the plaster to drain the melted wax, and another for pouring the resin. These holes can be made during the casting process with two greased, pointed sticks; or after casting, with a drill. Use care when melting the wax, because if it gets too hot the plaster will dry and crumble, or soldered connections on the stubs will melt.

base insulator

Forming the base insulator can be tricky, but here are some hints to make the process easier.

First, drill three holes in the ground-

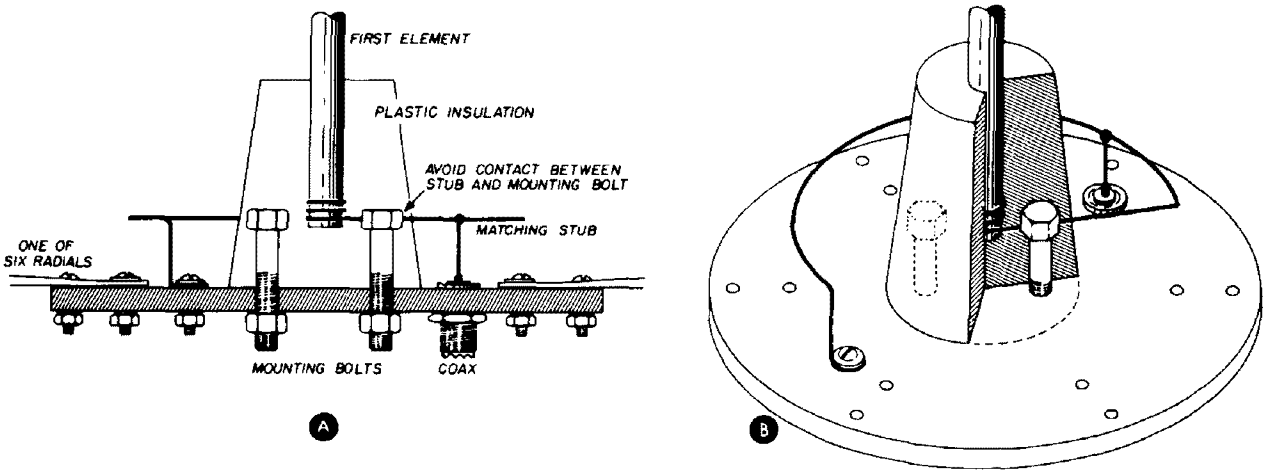


fig. 5. Detail of center insulator, bottom matching stub, and radial anchor plate, A. View B shows the assembly in perspective.

plane disc of the proper size to accept the bolts that will hold the antenna. Then drill another hole in the center of these three holes. (This hole will be used to pour paraffin into a mold.) Turn the ground-plane disc upside-down and place the three bolts into position, threading the nuts on the bolt ends to support the weight of the bolts as they hang (also upside-down) through the holes. See **fig. 7.**

Place the entire assembly on top of a paper cup, with the bolts hanging down inside the cup, but not touching its sides. The cup forms a mold for the paraffin, which is cast to form a pattern for the mold, as before.

After casting the paraffin, turn the base over and carefully remove the paper cup. Place a cardboard cylinder, or a can with the top and bottom removed, around the paraffin pattern to hold the plaster. Arrange a brace to hold the tubing vertically. Heat the end of the tubing, and let it melt its way into the wax pattern until the tubing reaches the proper position. (You'll probably lose some paraffin through the slot melted by the stub wire and have to retouch your pattern.)

When the pattern is ready, be sure that the base plate is absolutely level and the tubing is absolutely vertical. Then pour the plaster. Check to ensure that nothing has shifted out of alignment. When the plaster has solidified, you can melt the wax pattern, as before. It will probably be easier to do this without the radials installed.

casting in resin

With the molds ready, you can cast all the insulators in plastic at the same time, using only one batch of resin. This will eliminate considerable waste of resin and acetone. Use polyester or slow-curing epoxy resin, of the type used for fiber-glass boats, for the casting. Do not, under any circumstances, use the rapid-curing type of epoxy resin, which is commonly supplied with patching kits. The heat generated by the rapid curing causes a chain reaction inside the mold, because the heat can't escape as fast as it's generated. It can get hot enough to boil, which causes bubbles; or it can even catch fire. When the reaction is over and the

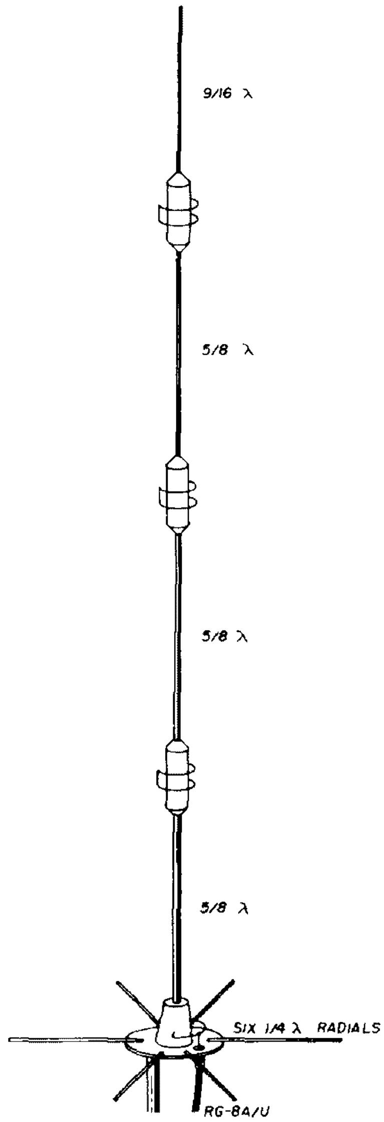


fig. 6. Overall view of the complete assembly.

stuff cools, it will crack. If polyester resin is used, apply twice the usual amount of hardener or catalyst. This will cause the mix to set up a bit faster and will also make the finished casting more flexible and less brittle.

Always measure carefully, and stir with care to avoid mixing bubbles into the plastic. Keep everything out of direct sunlight, and never mix more than you can use at one time. Otherwise the resin will gel, taking on the consistency of something between Jello and rubber. Note: the only thing that will remove resin from hands and tools is acetone. It evaporates so fast you have to see it to believe it, so get about twice as much as you think you'll need.

When the resin has set, you can break away the plaster molds, put everything together except the stubs, and proceed with tuning.

tuning

The stub wires should be soldered or welded to their radiating elements, but not yet connected to each other.

Prune the stub wires to exact length, using capacitive coupling – as loose as possible – to the hot end of a grid-dip oscillator tank coil. Using a receiver to check the oscillator frequency, resonate the three lower elements, with their stub wires, at one wavelength each. Ground one end of the top element to a metal screen or sheet and resonate it to ¾ wavelength. While doing this, try to keep all the stub wires parallel to maintain constant capacitance between them.

Install the radials and tune them all to ¼ wavelength. Now solder the wire ends together, or solder a shorting wire in place, to complete all the phasing stubs; then temporarily ground the end of the matching stub. Check the resonant frequency of the entire antenna with the

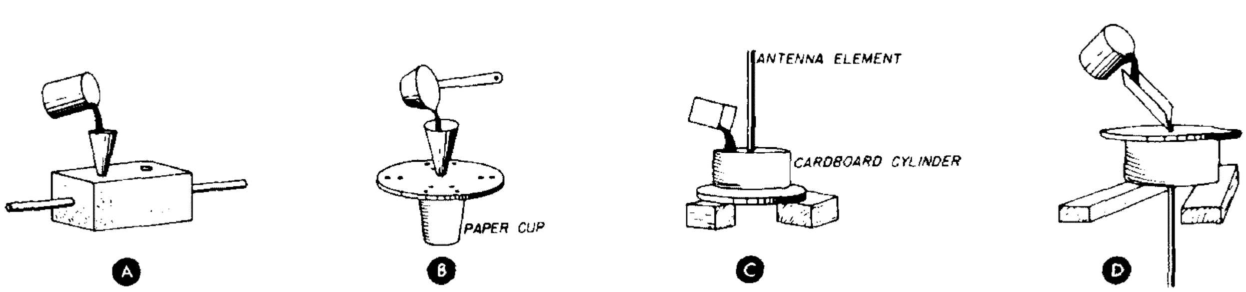
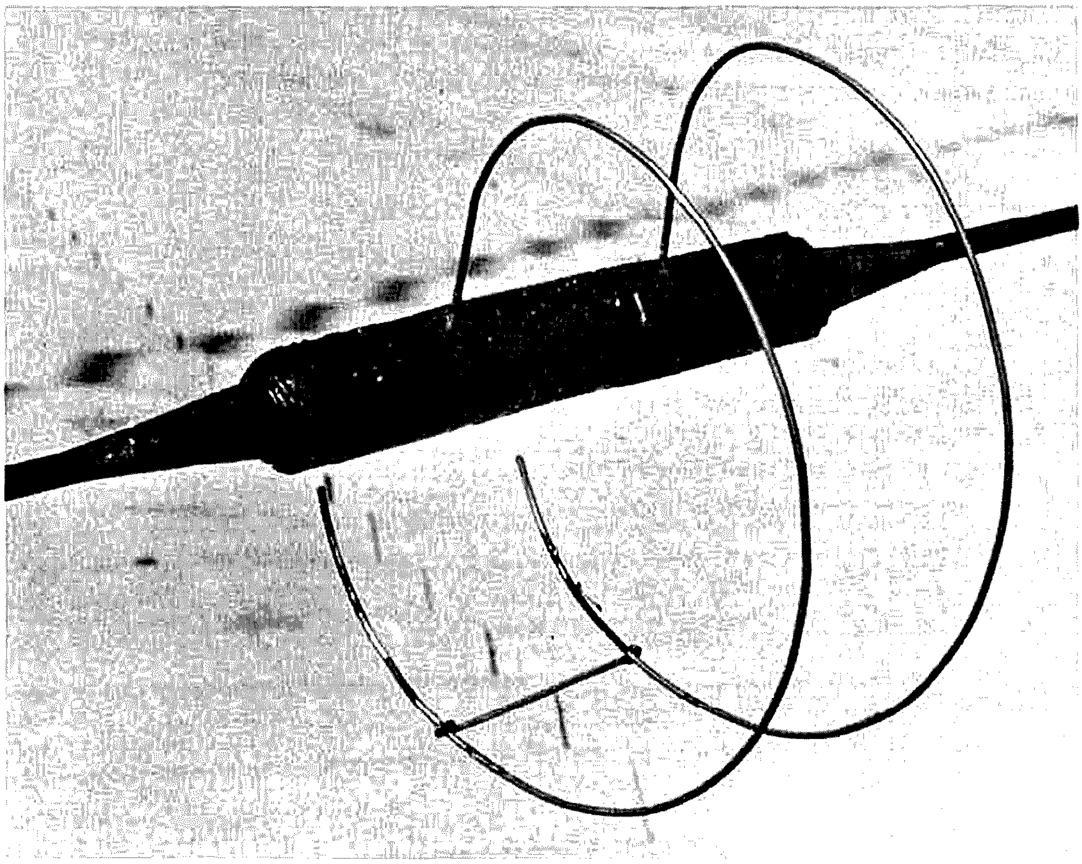


fig. 7. Making the insulators. Final step in casting an element insulator in resin is shown at A (see text) The base insulator is made by pouring a wax pattern, B, around which a plaster mold is formed, C. The wax pattern is then melted out of the mold, and the mold is filled with polyester resin, D.

gdo capacitively coupled to the top end of the antenna, as before. Note that the antenna will resonate at any odd multiple of the quantity (operating frequency divided by 19). For example, if the operating frequency is 190 MHz, the antenna will also resonate at 170, 150, 130, etc., and 210, 230, 250 MHz, etc. So be careful not to tune it on the wrong harmonic. With this in mind, adjust all the stubs equally, a little at a time if necessary, to get the antenna on frequency.

Now set up the antenna in a clear area, at a convenient working height for adjustment of the matching stub. Temporarily connect the coax to about the middle of the stub, and check the swr. If it isn't 1 to 1 (hi hi), move the coax a bit to either side and test again. When you've found the point of minimum swr, and if it isn't quite close to 1 to 1, then shorten the stub a bit by moving the ground point, and try again. If this doesn't improve things, try lengthening the stub. By alternately adjusting the length of the stub and the position of the tap, you should be able to get close to a 1 to 1 match. When you're satisfied with the swr, drill, punch and/or file a hole at the proper place and install a chassis-type coax connector. Reconnect the coax to the connector, and recheck the swr. You may have to readjust the tap a fraction of an inch.



Matching-stub detail. Stubs are tuned initially using a grid-dip oscillator, then pruned to correct length and soldered or welded to elements.

Now you can put 'er up as high as possible, and sit back and enjoy all the new signals you'll hear on the two-meter band.

references

1. *ARRL Handbook,* Headquarters Staff, American Radio Relay League, Newington, Connecticut.
2. R. L. Crawshaw, WA0NGV, "5/8 Wavelength Verticals," *73,* May, 1970, p. 36.

ham radio

adjustable balun for yagi antennas

Complete design and construction details for adjustable balanced-to-unbalanced feedline transformers

William I. Orr, W6SAI, EIMAC Division of Varian, San Carlos, California 94070

How convenient it would be if all Yagi beams presented a 50-ohm, unbalanced, non-reactive load at the feedpoint! Unfortunately, this is not the case, as most single-band Yagis exhibit a balanced feed-point impedance in the range of 18 to 30 ohms. Off-resonance, the impedance becomes complex, varying in the typical manner shown in **fig. 1A**. Moreover, when using a center-fed dipole the feedpoint impedance is balanced to ground, and an unbalanced coaxial feed system can result in degraded antenna performance. Poor front-to-back ratio, noise pickup and TVI may be some of the problems arising from improper attention to system balance.

A number of interesting matching and balancing systems have been evolved from time to time to solve these problems, and the better solutions work quite well. This article discusses an adjustable transformation linear balun of a type that has seen service in commercial installations over the years. However, its use in amateur circles has been restricted, possibly because of lack of knowledge regarding its operation. The balun will provide an adjustable step-up impedance match plus an accurate transformation from an unbalanced to a balanced mode for load values of about 10 to 50 ohms. Best of all, it is easy to adjust. Here's how you design, construct and tune this interesting device.

the L-network

The adjustable balun is derived from the basic L-network shown in fig. **1B**. By the use of a combination of a series and shunt reactance the low-impedance load may be matched to a high-impedance source. Transformation ratios up to ten or more are common. Two lumped constant L-networks are shown in fig. **2**. They are conjugate networks, the sign of the series and shunt impedances being reversed between the A and B versions. The A network is rather common in amateur equipment;[1] it may be recognized as the output section of the popular pi-L network circuit.

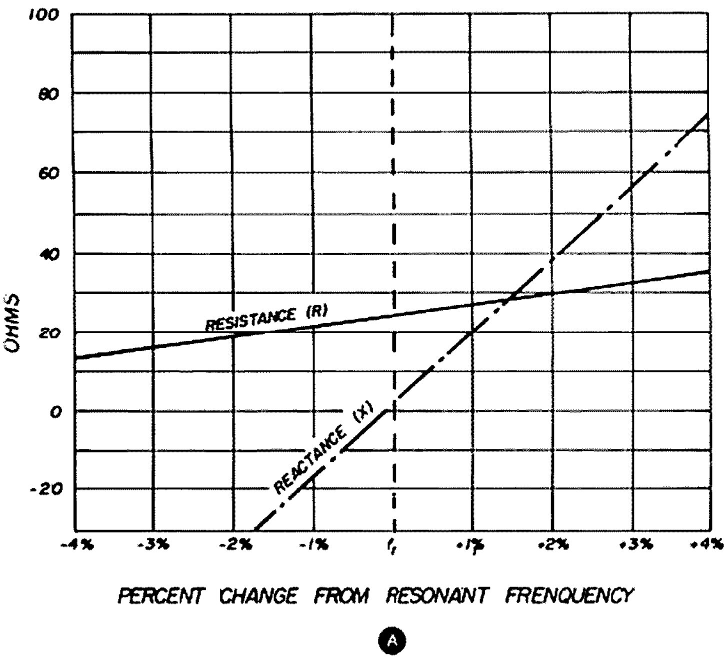
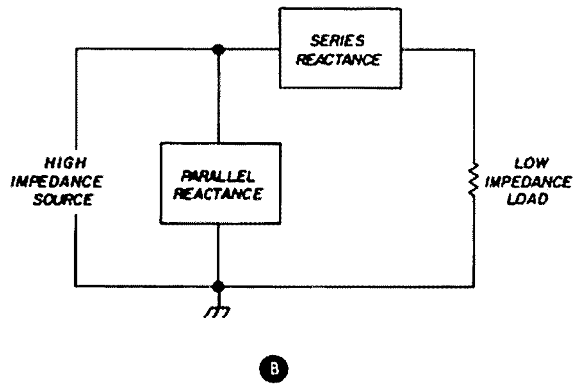
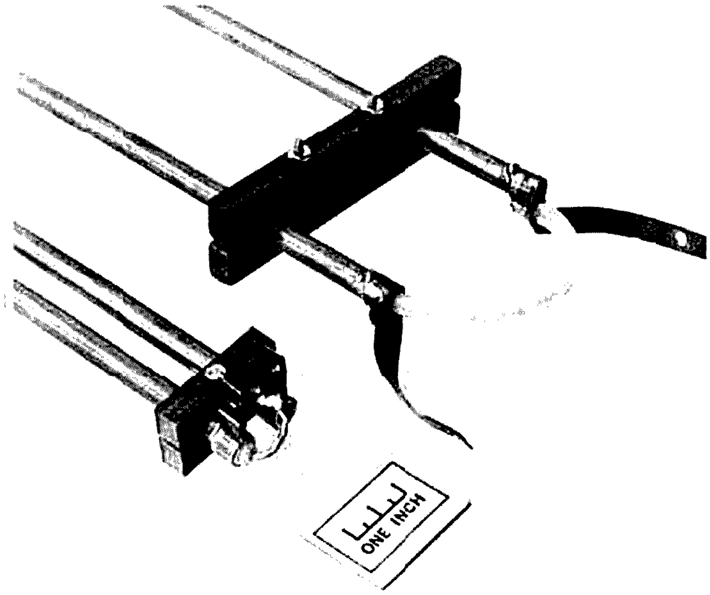


fig. 1. Yagi-beam impedance plot in A is similar to that of a dipole. Resistive component increases with frequency; reactive component is negative below resonant frequency and positive above it. L-network in B consists of series and shunt reactances and may be used to match a Yagi to 50-ohm coaxial line.



Vhf (left) and hf (right) balun construction. Balun tubes are locked in position by phenolic blocks. Inner conductor of coaxial line crosses over and is soldered to opposite tube. Short lengths of copper ribbon provide easy connection to dipole element.

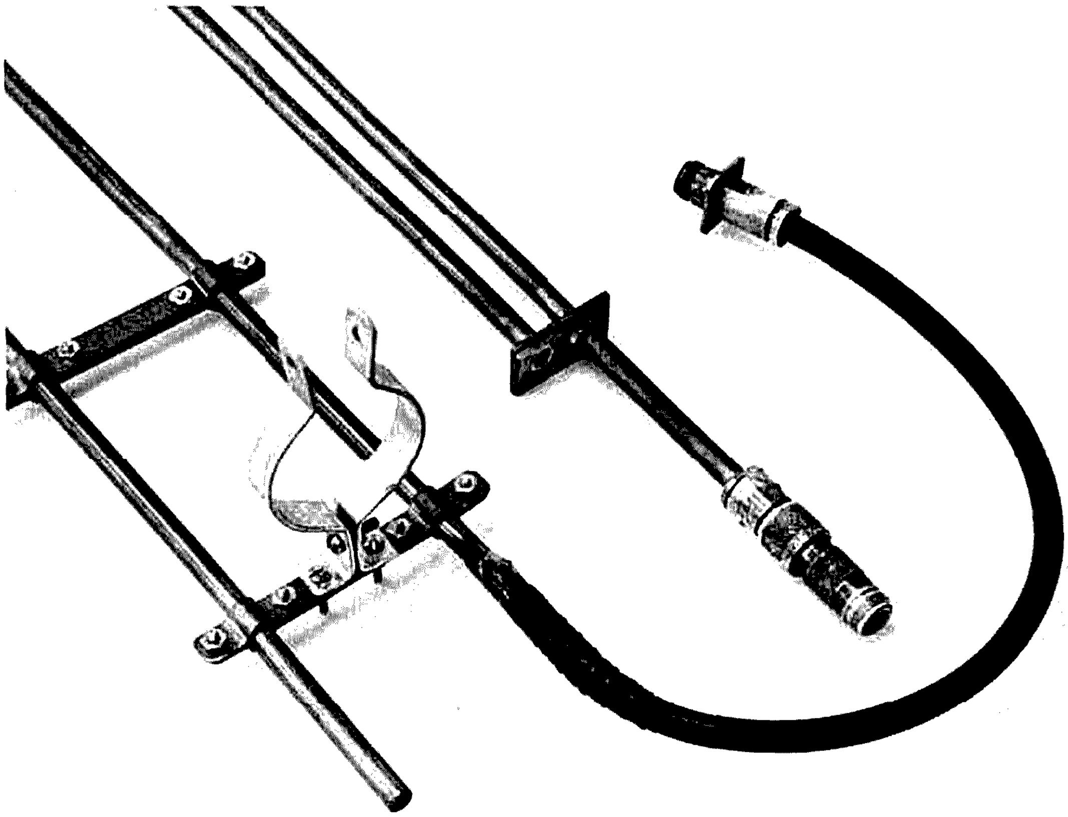


The configuration in fig. **2B** is used less often although it performs in the same general fashion as the A network. This second version of the L-network is the one used in the adjustable antenna balun described here.

In both versions, X_L and X_C represent real components, and R represents a resonant antenna load. If the antenna is deliberately made *non-resonant,* however, it may be adjusted to simulate a complex impedance containing the desired value of either X_L or X_C. Specifically, if the dipole antenna element is longer than resonance, it exhibits inductive reactance (X_L) at the center terminals; if it is shorter than resonance, it exhibits capacitive reactance (X_C). Thus, varying the length of the antenna beyond the resonant point eliminates the real components X_L (in network A) and X_C (in

network B). The L-network may be reduced to an off-resonant antenna, and either a parallel capacitor or inductor may be used, depending upon the type of network.

may take the form of a shorter-than-resonant dipole. In addition, the lumped inductor X_L may be removed and a shorted segment of transmission line substituted as shown in fig. 4.



Base end of hf (left) and vhf (right) baluns. Adjustable shorting bar is visible on hf balun. Extra shorting bar has aluminum boom clamp mounted on it. Coaxial line passes out of balun tube and has female type-N coaxial fitting (UG-23B/U) fitted at end of line. Vhf balun has one adjustable shorting bar (not in photograph) with copper shorting bar sweated to bottom of balun tubes. One tube is a few inches longer than the other and has a modified UG-18B/U coaxial plug soldered to the tube. In the hf balun, the coax braid is soldered to the end of the copper tube and the joint wrapped with vinyl electrical tape. The braid is completely removed in the vhf balun with the inner conductor of the line soldered directly to the coaxial fitting; the inner conductor is passed through the tube to the opposite end.

balanced L-network

There still remains the problem of connecting an unbalanced coaxial transmission line to a balanced antenna element. Fig. 3 shows the network of fig. 2B redrawn for a balanced condition. The ground point is moved to the center of coil X_L, and two capacitors, each double the value of X_C are placed in series with the load. As before, if the load is considered to be an antenna, the series capacitance

The configuration is the mechanical heart of the adjustable balun. A shorted transmission line less than one-quarter wavelength long presents an inductive reactance at its open end. A shorter-than-resonance dipole presents a capacitive reactance at its terminals. If the length of the line segment (or stub) and the length of the dipole are properly chosen, the components of the impedance matching network are reduced to a few lengths of

tubing, two of which are the halves of the driven element in the beam antenna.

The linear L-network is easily converted to a practical balun transformer as shown in fig. 5. Points A and B of the balun are balanced to ground and present the correct impedance to match the shortened dipole. The unbalanced coaxial line may be brought into the balun through one of the balun tubes, with the center conductor of the coaxial line crossing over at the end of the device to contact the opposite balun tube, as shown in the illustration. Of course, the impedance of the coaxial line must be held to the original value as it passes through the balun tube. By adjusting both the shorting bar on the balun and the length of the dipole this simple device will provide excellent balance and transformer action.

Balance is achieved by permitting the outer shield of the coaxial line to assume the potential of the balun tube as it passes from the grounded end (C) to the terminal end (B). Cross-connecting the center conductor to the opposite balun leg insures 180° phase reversal is maintained.

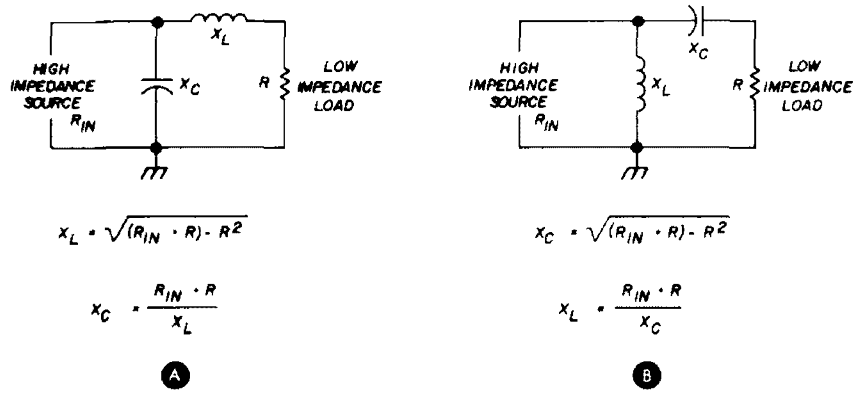


fig. 2. Conjugate L-networks. Network A is preferred when load reactance may be negative. Network B is preferred when load reactance is positive. For reactance ranges common to Yagi antennas, either network may be used.

network transformation

The reason the L-network is able to transform one impedance value to another is that, for any series circuit consisting of a series reactance and resistance, an equivalent parallel network which possesses the same impedance characteristic can be found.[2] The general case for determining the values required for any two impedances to be matched by the L-networks of fig. 2 are summarized by the following equations:

$$\frac{R_{in}}{R} = Q^2 + 1 \qquad (1)$$

$$Q = \frac{X_s}{R} \qquad (2)$$

$$Q = \frac{R_{in}}{X_p} \qquad (3)$$

Where R_{in} = the input impedance, R = the load impedance, X_s = the series reactance, X_p = the parallel reactance, Q = the circuit Q, and the series and shunt reactances are of opposite sign.

For easier usage with 50-ohm lines, these formulas may be reduced to the ones shown in fig. 2 with the relationship

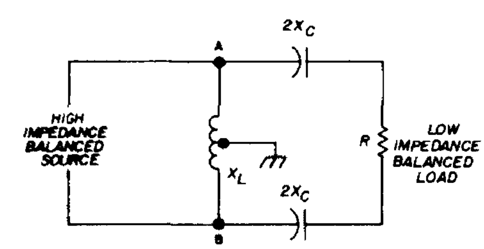


fig. 3. The L-network of fig. 2B redrawn for a balanced source and load. Points A and B are at equal and opposite potential to ground.

between X_s, X_p, R_{in} and R given in **fig. 6.** The reactance values are in ohms and may be translated to picofarads and microhenries with the aid of a reactance chart.*

balun design

Once the transformation ratio and the values of series and parallel reactance have been established, the physical balun may be designed from transmission-line formulas. The fact that a shorted, two-conductor transmission line of the proper length exhibits inductive reactance at the terminals makes it possible to substitute such a line for the inductor in an L network. The amount of reactance shown by the line segment is determined by the characteristic impedance and the electrical length of the two-conductor line. The inductive reactance of a shorted lossless balun line, less than a quarter-wavelength long, constructed in this fashion, is:

$$X_L = Z_o \tan l$$

where X_L = inductive reactance in ohms, Z_o = characteristic impedance of the balun line, and l = length of the balun line in electrical degrees.

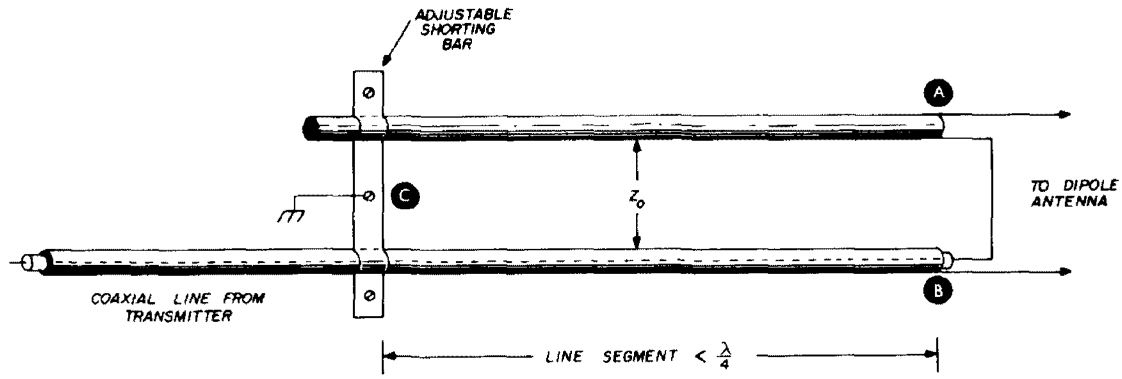


fig. 5. Linear transformer of fig. 4 is modified into balun by passing coaxial line down one leg. Points A and B are balanced to ground. Inner conductor of coaxial line is cross-connected to opposite balun leg. The impedance transformation is adjusted by varying length of the balun and length of driven element of antenna.

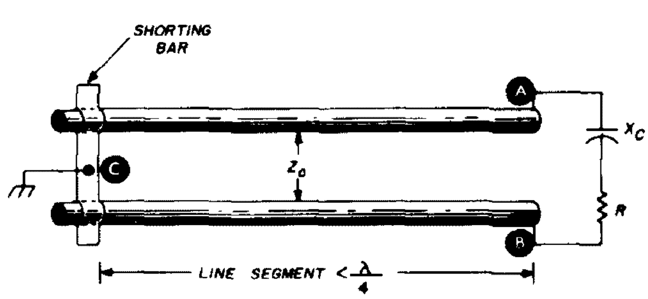


fig. 4. The inductor X_L in fig. 3 may be replaced with a segment of shorted transmission line. Points A and B are at equal and opposite potential to ground, C.

Fig. 7 is a plot of balun line length (l) in electrical degrees as a function of the ratio of load impedance to balun impedance (σ). A plot of the ratio σ in terms of line length in feet for the 20-meter band is given in **fig. 8.** These charts provide sufficient information to build your own linear balun.

practical balun transformer

A balun transformer built along these principles is shown in the photographs. For convenience the balun is made of 3/8-inch diameter hard-drawn copper tubing. The feedline, RG-8A/U cable, will just pass through the tubing when the braid and the vinyl jacket are removed from the line. Using a center-to-center spacing of 3 inches, the balun line will have a characteristic impedance (Z_o) of about 325 ohms.

The Smith chart may also be used for impedance transformation. See P. H. Smith's book, "Electronic Applications of the Smith Chart," published by McGraw-Hill.

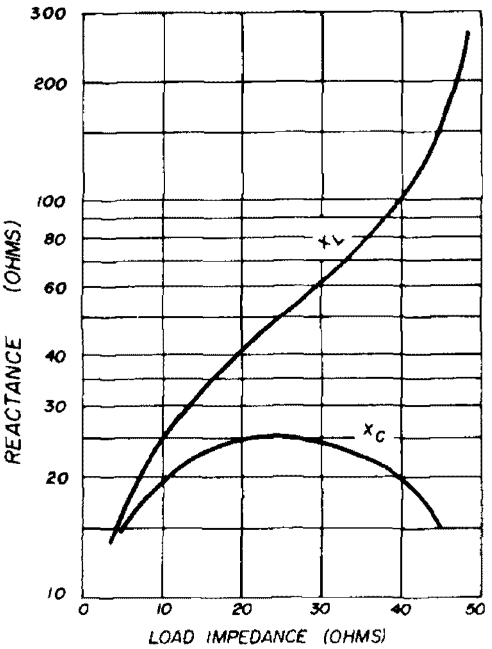


fig. 6. X_L and X_C values for antenna load R. This chart may be used for determining balun reactance values when a 50-ohm transmission line is used. For example, if the load impedance (antenna impedance at resonance) is 25 ohms, the capacitive element (X_C) of the balun of fig. 2B is 25 ohms, and the inductive element (X_L) is 50 ohms. The chart may be used with the balun of fig. 2A if the nomenclature of the curves is reversed (X_L becoming X_C and X_C becoming X_L).

Assume the impedance of the transmission line is 50 ohms and the antenna load is 20 ohms. Using fig. **6**, the value of X_C is found to be −24.5 ohms, and the value of X_L is +41.5 ohms. Turning to fig. **7**, the ratio of X_L to balun characteristic impedance (σ) is $41.5/325 \approx 0.127$ as noted on the y-axis. The value for *l*, as found on the x-axis is about 7.5 electrical degrees.

To get the answer directly in feet, fig. **8** may be used for the 20-meter band. In this example for $\sigma = 0.127$ (read as 0.125 on y-axis) the balun length is about 1.4 feet, or 16 inches.

The chart of fig. **6** has indicated that the series reactance (X_C) for this example is −24.5 ohms. This reactance takes the form of a shorter-than-resonance driven element. The amount of shortening required is a function of the length to diameter of the element and the feedpoint impedance at resonance of the element.* The amount of shortening may be computed easily for a single dipole element, but no information exists (that I am aware of) that permits this computation to be made for a multi-element Yagi beam. Consequently, the shortening necessary to bring the driven element to the proper reactance value is best determined by the heuristic method — cut and try! For a three-element 20-meter beam, shortening the driven element about three to six inches each side seems to bring things into the ball-park.

*As antenna length decreases, feedpoint impedance decreases, too. The decrease typically runs about 10% to 15% for the range encountered in matching a three-element Yagi beam to a 50-ohm transmission line.

adjusting the balun

The balun transformer may be pre-set and attached to the beam antenna. The

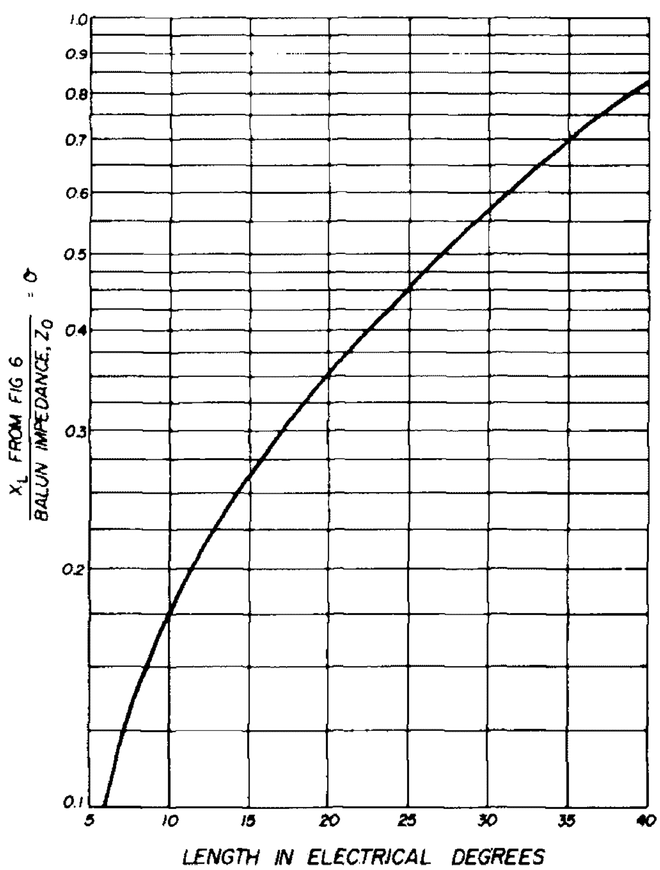


fig. 7. Balun length in electrical degrees as a function of the ratio of the load impedance to the balun impedance (X_L/Z_0).

balun can run parallel to the boom for convenience at a distance of about six inches from the boom. Positioning the balun closer to the boom will necessitate a change in setting. The driven element, for a starter, should be shortened about three inches on each tip (for 20 meters).

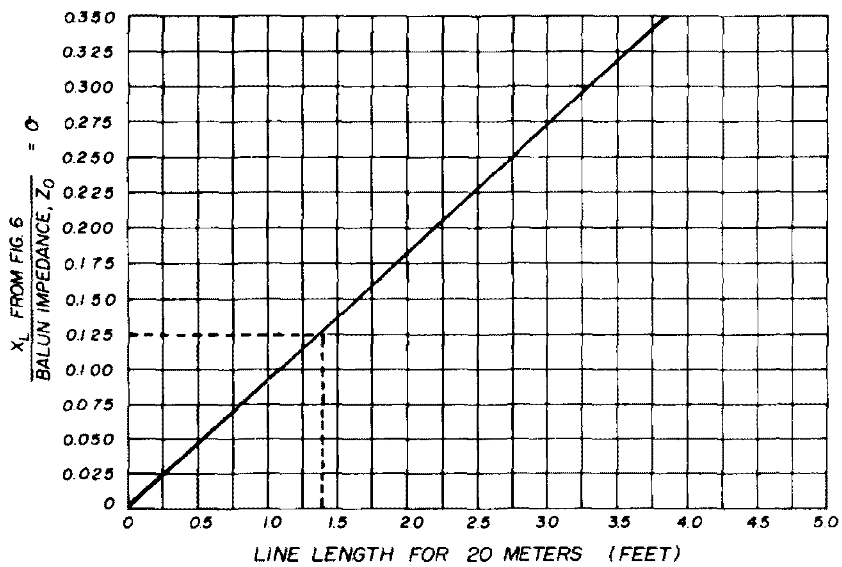
The excellence of adjustment is ascertained by running an swr curve across the band, making a measurement every 50 kHz or so. Balun length and driven-element length are then adjusted to drop the swr curve to a 1-to-1 ratio at or near the center of the band. Adjustment is not critical, and if you log your adjustments you will quickly be able to estimate the degree of change necessary to adjust the system "on the nose." Adjustments of balun and dipole length are interlocking, but setting the balun to the length in **fig. 8** for a given value of antenna load and preshortening the driven element a few inches will insure that the starting point is not too far out of line.

For convenience the feedpoint impedance of the three-element Yagi beam may be taken as 20 ohms. In fact, it is possible to calibrate balun length versus terminal impedance in the home workshop using a grid-dip oscillator, an antennascope or swr meter and a handful of 1-watt composition resistors of known value. The calibrated transformer balun may then be used backwards, as it were, to determine the feedpoint impedance at resonance of the antenna.

Various adjustable baluns are in use at W6SAI. A permanent one is placed on the beam antenna, and two others are calibrated for use around the shack on experimental antennas. The balun for high frequency work is about five-feet long and has center-to-center spacing of 3 inches. The vhf balun is about the same length (approximately 1½ wavelengths at 144 MHz) with a center-to-center spacing of 1¼ inches. The extra half-wavelength was added to the vhf balun to permit measurements to be made without the operator being in the immediate field of the antenna.

Baluns of this general type may also be used as step-down transformers to match balanced load impedance in the range of 50 to 300 ohms to low-impedance coaxial lines — but that's another story.

fig. 8. Balun conversion chart for 20 meters. Balun length in feet may be determined if resonant antenna load and balun impedance are known. Ratio of these two items is found on y-axis, and balun length is read on x-axis. Chart may be used for other bands (multiply lengths by 2 for 40 meters, divide by 1.5 for 15 meters, divide by 2 for 10 meters, etc.



references

1. William I. Orr, W6SAI, editor, "Radio Handbook," 18th edition, p. 288, Editors & Engineers, New Augusta, Indiana.
2. Gray and Graham, "Radio Transmitters," Chapter 5, Mc-Graw-Hill, New York,

ham radio

a practical experimenter's approach to time-domain reflectometry

Time-domain reflectometry represents one of the easiest methods of analyzing transmission-line problems. The system shown here is easily duplicated in your shop

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The use of time-domain reflectometry for analyzing transmission lines has enjoyed little popularity among experimenters who must finance their own projects. In fact, there are probably many amateurs who have never heard of this technique. Essentially, time-domain reflectometry is a closed-circuit radar system that displays transmission-line impedance "bumps" and discontinuities on an oscilloscope.*

This system has been used by repair crews for many years to locate faults in high-voltage transmission lines. A pulse burst is sent continuously down the transmission line. If the pulse encounters a short or open circuit the reflection travels back to the sending point, where it is compared in phase, time and amplitude with the original pulse. This comparison indicates the distance to the fault as well as its nature.

With new high-speed oscilloscopes it has become practical to apply the time-domain reflectometry technique to high-frequency transmission lines. The faster the transmitted pulse, the greater the distance resolution, since distance is related to time. Hence, distance resolution has shrunk from hundreds of yards to fractions on an inch, and commercial systems permit accurate measurement of reflections only a thousandth of a volt in amplitude.

Although commercial time-domain reflectometry equipment costs in the thousands of dollars, the basic technique is quite simple, and if the experimenter is willing to sacrifice the extreme accuracy of commercial equipment, he can duplicate the basic system at relatively low cost.

The only expensive requirement is a good oscilloscope, preferably one with triggered sweep, with bandwidth extending to 10 MHz or beyond. The upper limit on scope bandwidth is the prime limiting factor in measuring small discontinuities along a transmission line. Aside from the requirement for a good scope everything else can be built or is readily available in your shop. With this

*For more information on time-domain reflectometry, see Hewlett-Packard application note 67, "Cable Testing with Time-Domain Reflectometry." Available from Hewlett-Packard, 1501 Page Mill Road, Palo Alto, California 94304.

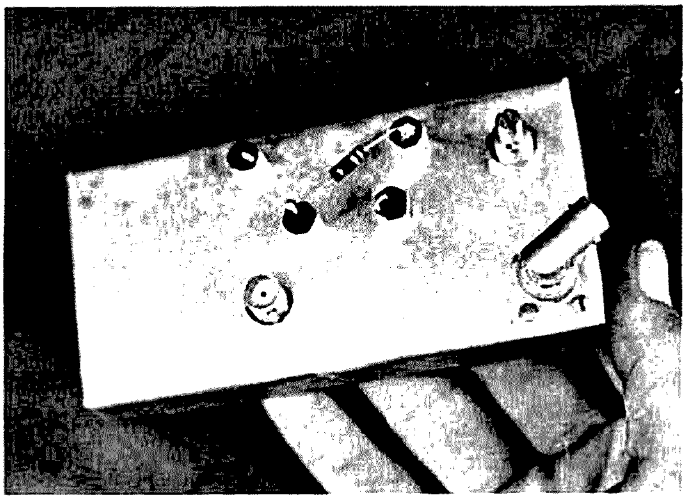
simplified approach transmission-line impedance cannot be measured within a tenth of an ohm accuracy, nor can transmission-line lengths be measured in one-foot lengths, but accuracy and resolution are good enough to be quite useful, and will provide answers to transmission-line problems where none were previously available.

fundamentals

The time-domain reflectometry (TDR) system is based on the propagation of an energy pulse down a transmission line into the load at the other end. A scope is used to monitor the pulse. It takes a finite length of time for the pulse to travel down the transmission line; this length of time is very short and depends upon the length of the transmission line. As the pulse travels through the line it may or may not be upset. If discontinuities exist along the line, part of the pulse energy will be reflected back toward the generator.

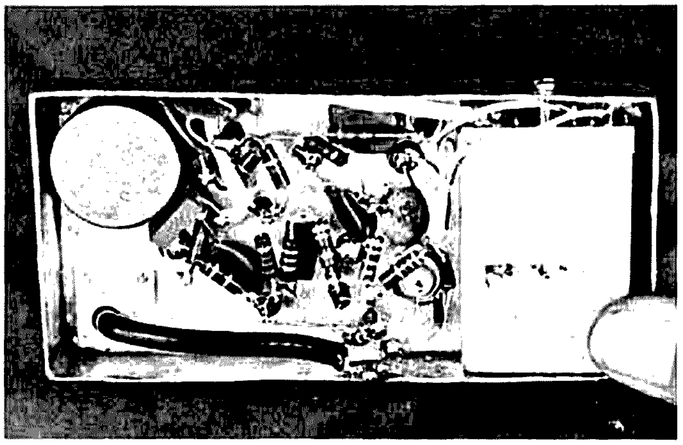
Since the scope is connected across the output of the pulse generator any energy that is reflected will be displayed. The original outgoing pulse is sharp-cornered and flat-topped; when reflected energy is added the flat top develops ripples and humps. The position of the ripple along the normally smooth top of the pulse corresponds directly to the length of time taken for the pulse to travel to the discontinuity and back again.

Schmitt-trigger pulse generator and 1-MHz source are built into small steel chassis.



If the transmission line is perfectly matched the pulse will be loaded down, and its voltage amplitude will be reduced, but the pulse top will remain smooth and flat. On the other hand, if the coax has water in it or is otherwise contaminated, or if a splice is poorly done (or any number of similar maladies) there is a mismatch and the pulse displayed on the face of the scope will be distorted. The oscillographs, figs. 7 through 10, show typical forms of distortion.

The input pulse must be very short and fast, and the scope must have charac-



Inside the time-domain reflectometry unit. The 1-MHz source is in the shield can to the left.

teristics equal to (and preferably better than) the pulse's fast rise time and short duration. The duration of the pulse determines the distance resolution of the system. For example, a two-second pulse is short enough for a transmission line 186,000 miles long (the pulse must travel twice the length of the transmission line before it gets back to the scope, thus, two seconds instead of one). Although the velocity factor of the coax, typically 0.66, will slow things down somewhat, pulse lengths this long are obviously not practical. However, the pulse can be easily shortened, and a pulse length of 1 microsecond is suitable for approximately 324 feet of coax or 480 feet of open-wire line. This puts transmission-line lengths within manageable limits.

In commercial time-domain reflectometry systems much shorter pulses are

used with special sampling oscilloscopes. In a typical setup the combination of a very-fast-rise-time pulse generator and sampling scope provides an overall frequency response of 2.3 GHz or better.

equipment

Figs. 1 and **2** show two different approaches to getting a workable time-domain reflectometry system going. **Fig. 1** was my first attempt; nothing was built

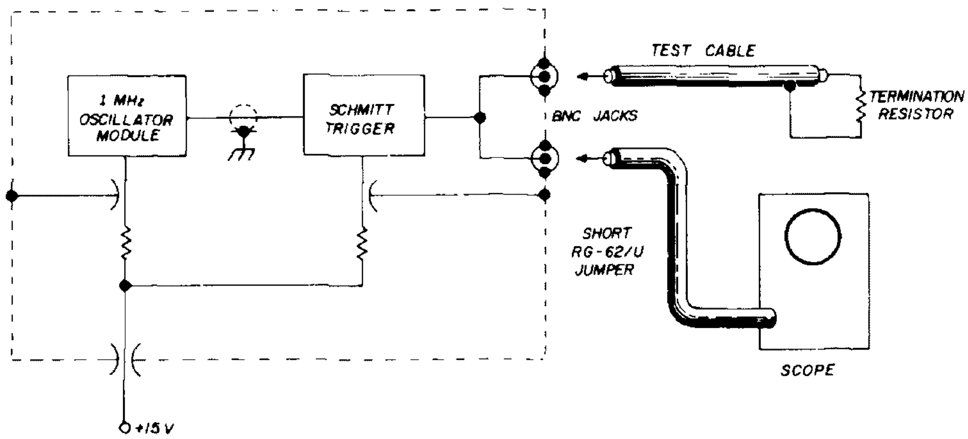


fig. 2. Practical time-domain reflectometry system uses 1-MHz crystal-controlled oscillator, Schmitt-trigger pulse generator and high-frequency oscilloscope.

This corresponds to a rise time of 150 picoseconds (150 trillionths of a second) which means that transmission-line discontinuities less than a half-inch apart may be resolved and isolated.

Although the pulse length for a practical amateur system doesn't have to be this short, the requirements for the pulse are fairly stringent. In my TDR system the pulse repetition rate was chosen at 1 MHz. For a fast-rise-time smooth pulse harmonics beyond 30 MHz will be present. Fortunately, adequate semiconductors are readily available, and the necessary construction techniques are no more than those required for a good 100-kHz crystal calibrator.

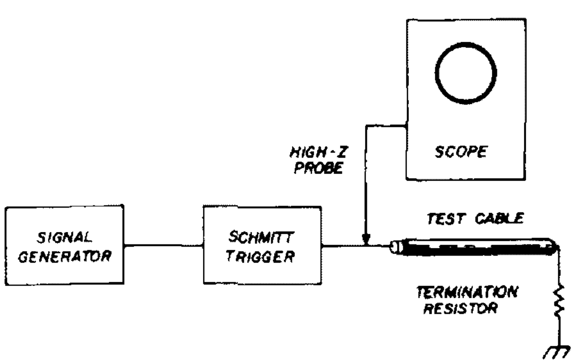


fig. 1. Original time-domain reflectometry system built by author indicated feasibility of simplified approach.

but the Schmitt trigger circuit, and it was only loosely haywired together. Friends viewed the assemblage with skepticism; the Schmitt trigger resembled floor sweepings more than anything else. However, it worked, and from that initial success evolved the system in **fig. 2.** This is basically the same except that a 1-MHz driving source was used. A crystal-controlled 1-MHz driving source is particularly valuable where sweep-time calibration of the scope is in question, since the pulse generator then doubles as a very accurate time-base calibrator.

The 1 MHz source I used was taken from commercially built equipment and puts out a crystal-controlled 1-MHz sine wave. It can be easily duplicated, or a standard signal generator may be used. It does not take much power to drive the Schmitt trigger. A simple 1-MHz driver circuit is shown in **fig. 4.**

The Schmitt trigger circuit is a standard design; component values were juggled in the prototype to provide a clean, stable pulse with fairly low impedance. The hysteresis of the circuit was made as low as practicable while still allowing sufficient pulse-length adjustment. The 2500-ohm pot sets the switching threshold, and the pulse length with a sine-wave driving signal. Pulse length is also a

function of the amplitude of the 1-MHz source. Resistor R1 is used to set drive-signal amplitude by loading generator output down to the level required by the Schmitt circuit.

sistor and the saturation voltage of Q2. When these voltages are offset pulse output will be referenced against a 0-volt baseline instead of being elevated above ground. The diodes are silicon switching

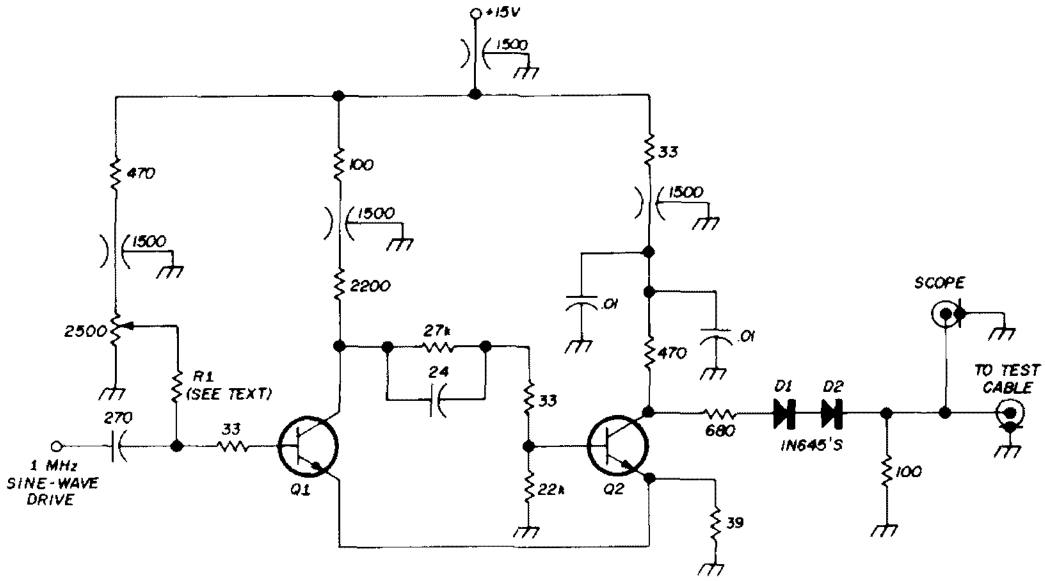


fig. 3. Schmitt-trigger pulse-generator circuit. Transistors are surplus silicon npn types; Motorola HEP50 are suitable. Value of R1 depends upon 1-MHz amplitude and source impedance.

The transistors are small-signal, high-frequency switching types. In my unit I used devices from surplus digital equipment. I don't know anything about them except that they are silicon npn. With the 2500-ohm threshold pot in the circuit transistor substitution should cause no problem, only readjustment of the pot.

When the final trigger circuit was built it worked when power was first applied, and was tested by hooking it into a communications receiver. As the threshold pot was varied for symmetrical on-off characteristics, the even harmonics went through a null of several dB as read on the S-meter. Since theory says that there will be no even harmonic energy in a perfect square wave it was assumed that the pulse generator was working properly. This was verified later with an oscilloscope.

The 680-ohm resistor in series with the Schmitt circuit output prevents excessive loading of the circuit due to low cable impedance. The two diodes in series with the output are to offset the voltage developed across the 39-ohm emitter re-

types and are not at all critical. I used 1N645s.

Adjustment of the 2500-ohm threshold pot is not critical. If the driving voltage at the base of Q1 has the proper amplitude the threshold action of the pot will occur across a few degrees of rotation near the center of its range. Once the

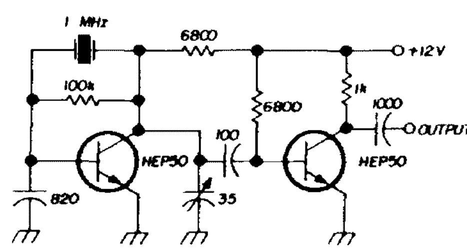


fig. 4. Crystal-controlled 1-MHz oscillator.

circuit is switching properly the narrow range of the pot's adjustment is used to vary the length of the positive pulse output. In my unit, 0.6 microsecond pulses provided ample time for work.

operation

The pulse generator output must be

hooked directly to the transmission line without excessive lead lengths. The scope connection is very critical. A high-impedance scope probe with a 10:1 divider is satisfactory, but the scope must have adequate sensitivity. I found it more expedient to hook the pulse output directly into the scope with a *very* short length (less than 6 inches) of low-capacity coax cable such as RG-62/U. If there are any ground loops present in the system they will probably show up as excessive overshoot and/or ringing on the front edge of the pulse. This is where high-frequency construction techniques are very important.

Once a clean pulse is visible on the scope (similar to fig. 5) scope sweep time can be adjusted to fill the entire screen with the top of the pulse as shown in fig. 6. At this point the pulse generator output is terminated directly at its output. The next step is to hook the unit to a transmission line. However, the following precautions must be observed: the transmission line must be terminated by a resistive load that is dc coupled, and there must be no frequency-selective devices (coaxial baluns, matching stubs, etc.) anywhere in the line. Impedance variations and discontinuities should now be visible on the top of the pulse; figs. 7, 8,

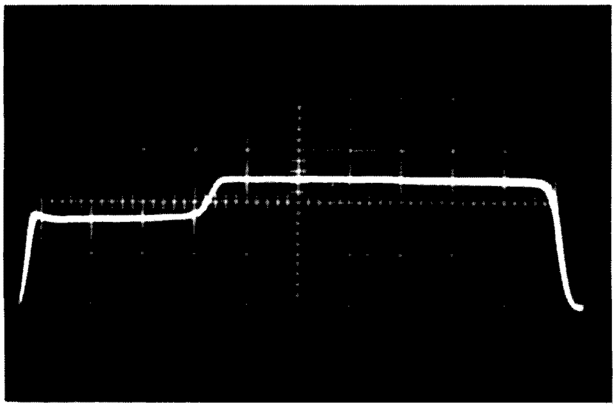


fig. 7. Transmission-line display. From left to right: 68-feet 50-ohm RG-8/U, 25-feet 93-ohm RG-62/U, 100-ohm termination. Termination occurs approximately at center screen; right half of pulse displays no information. "Step" in display occurs at transition from 50- to 100-ohm transmission line.

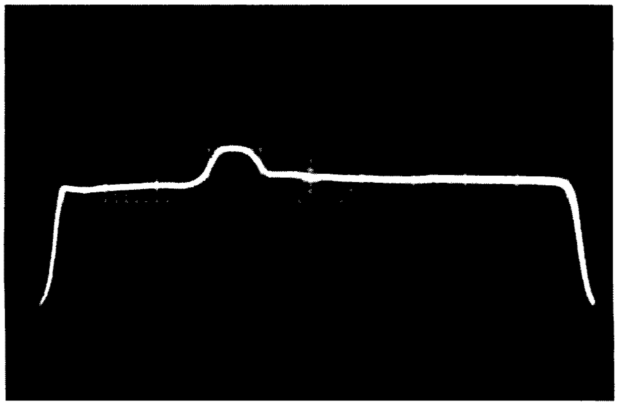


fig. 8. Transmission-line display. From left to right: 68-feet 50-ohm RG-8/U, 25-feet 93-ohm RG-62/U, 25-feet 50-ohm RG-58/U, 51-ohm termination. Bump indicates length of 93-ohm RG-62/U.

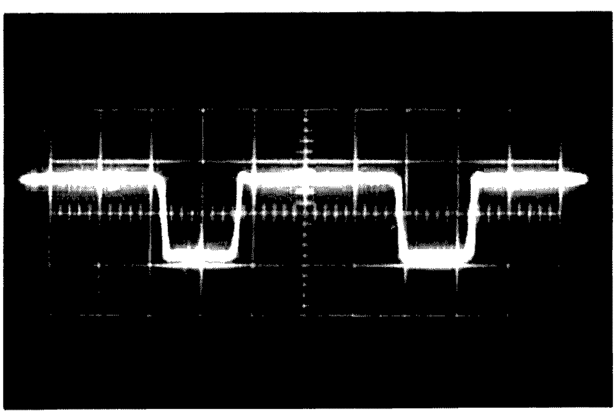


fig. 5. Rectangular pulse output from Schmitt trigger. Pulse rate, 1 MHz; sweep time, 0.2 usec per division.

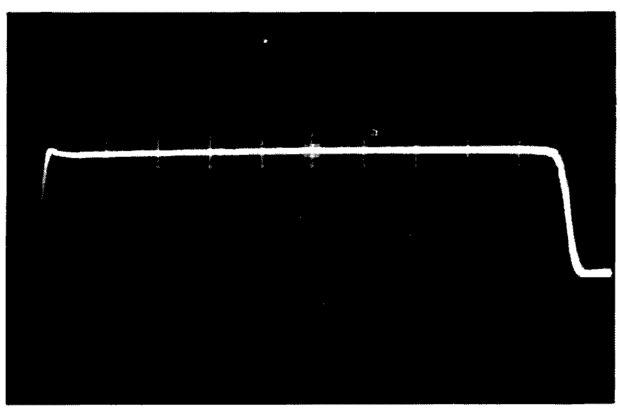


fig. 6. Same pulse as in fig. 5 but with sweep time decreased so top of one pulse fills scope face.

9 and 10 show examples of what to look for.

The scope display in fig. 7 shows the effect of increasing the impedance of the line. This trace shows a length of 50-ohm

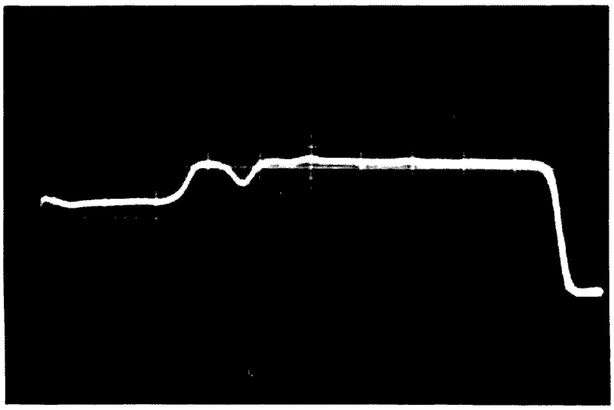


fig. 9. Transmission-line display. From left to right: 68-feet 50-ohm RG-8/U, 25-feet 93-ohm RG-62/U, 5-feet 50-ohm RG-58/U, 100-ohm termination. Because of scope's bandwidth limitation, the 5-foot section of 50-ohm line did not get down completely to the 50-ohm level.

line (RG-8/U) connected to a length of 93-ohm line (RG-62/U) followed by a 100-ohm termination. The step in the trace occurs at the connection point between the 50- and 93-ohm lines. The termination is at approximately the center of the screen; the right half of the trace contains no information.

In fig. 8 you can see another example of impedance increase. In this case three sections of coaxial line were connected together: a 50-ohm section followed by a 93-ohm section followed by another 50-ohm section and a 51-ohm resistor termination. The bump in the trace corresponds to the short section of 93-ohm line.

Fig. 9 shows a 50-ohm line connected to a 93-ohm line, followed by a short 5-foot section of 50-ohm line and a 100-ohm termination. The first step in the trace is the input to the 93-ohm coax. The dip to the right of the step indicates the location of the short section of 50-ohm line. Because of bandwidth limitations imposed by the oscilloscope, the pulse reflection from the 5-foot section of 50-ohm line did get down completely to the 50-ohm level.

In fig. 10 you can see the effect of a discontinuity in the transmission line. In this case 130-feet of 50-ohm coax was terminated with a 50-ohm dummy load. The discontinuity, a 24-pF capacitor shunted across the line, is indicated by the bump in the trace. The capacitor is 70-feet from the input. In this display the vertical gain of the scope was increased to make it easier to see the discontinuity. Note that overshoot at the beginning of the trace is now objectionable.

limitations

Since the output pulse has frequency components to 30 MHz or beyond, but no higher than 100 MHz, this low-cost system cannot be used for waveguide, Goubau-line or other transmission lines that have a low-frequency cutoff point. Although the system described here is designed for coaxial transmission lines

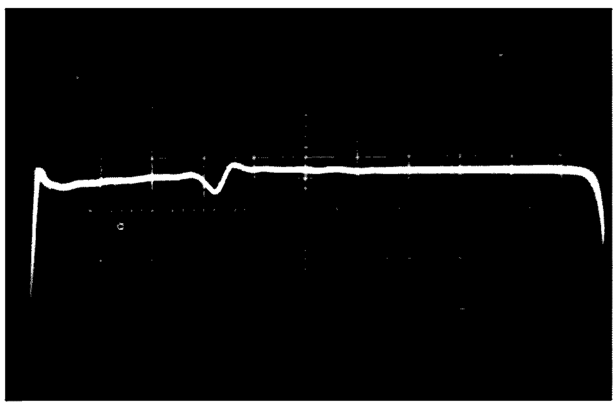
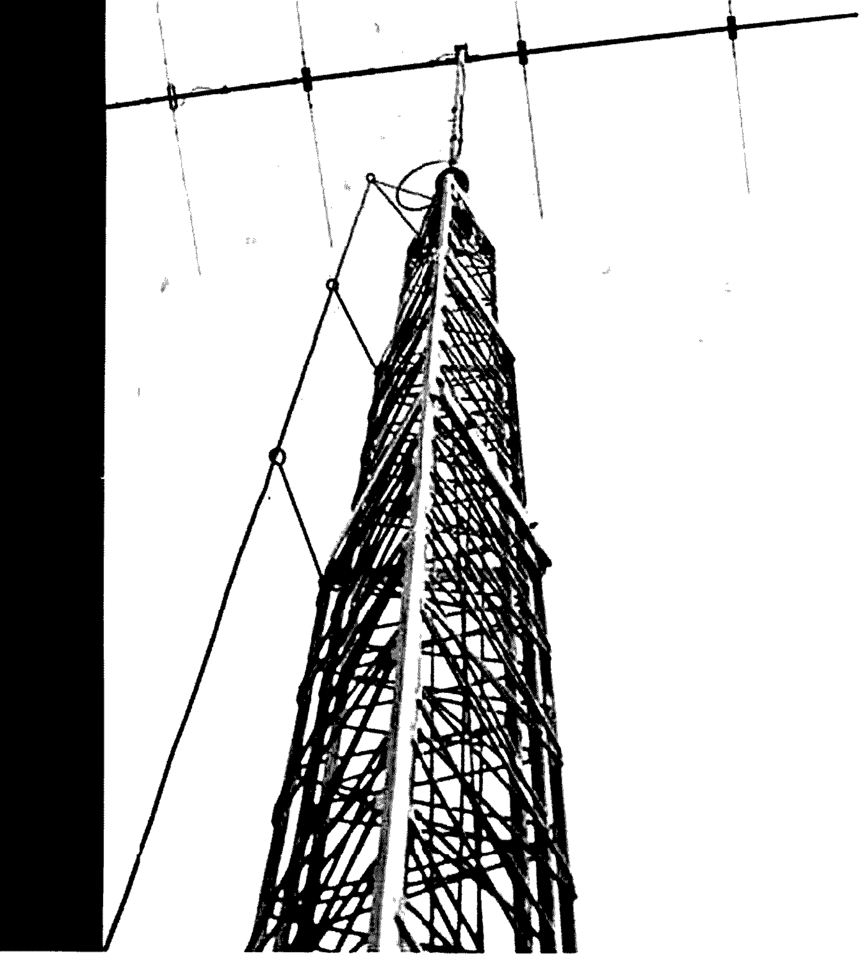


fig. 10. 130-feet of 50-ohm transmission line with 50-ohm termination. Discontinuity (24 pF shunted across line) at end of 70 feet. Vertical gain was increased to make the discontinuity more visible; note that overshoot is now objectionable.

there is no reason why a similar system with balanced output could not be designed for open wire line and twisted pair. This system provides tangible answers to questions that have previously been guessed at, and provides visible goals in otherwise hit-or-miss situations.

ham radio



homemade tilt-over antenna tower

Construction details for a triangular, 70-foot tower you can build for a fraction of the cost of a ready-made structure

Gene Nelson, WA3EWH, Belle Vernon, Pennsylvania

Many hams would like to have a good tilt-over antenna tower for as little money as possible. This was my wish also, but I didn't do anything about it until a good friend, WA3AHM, started needling me about building one. This article describes the result of my answer to this challenge – a 70-foot tilt-over tower for only \$44.62 in material costs. I used material purchased from local junk dealers for all structural members except the 3/8-inch round-rod diagonal trusses (fig. 1), which were donated by friends. If purchased, the 3/8-inch round rod would come to about \$15.00 extra – still a pretty good bargain when you consider the cost of a commercially built tower of this type. Accessories such as gears, lift motor, cables, and tilt-over winch are additional expense items, of course.

construction

This article is presented with the assumption that you've had some welding experience. If you haven't and wish to build the tower, you can have the entire assembly welded professionally and still come out ahead of the current market price for a tower of this size.*

I've included some hints based on my experience that should help if you've never undertaken a project such as this. The main point to bear in mind is that the verticality of the finished structure will depend on how accurately you measure the templates and the care with which the parts are positioned during welding. A small shift in alignment during assembly will be magnified many times in the completed tower.

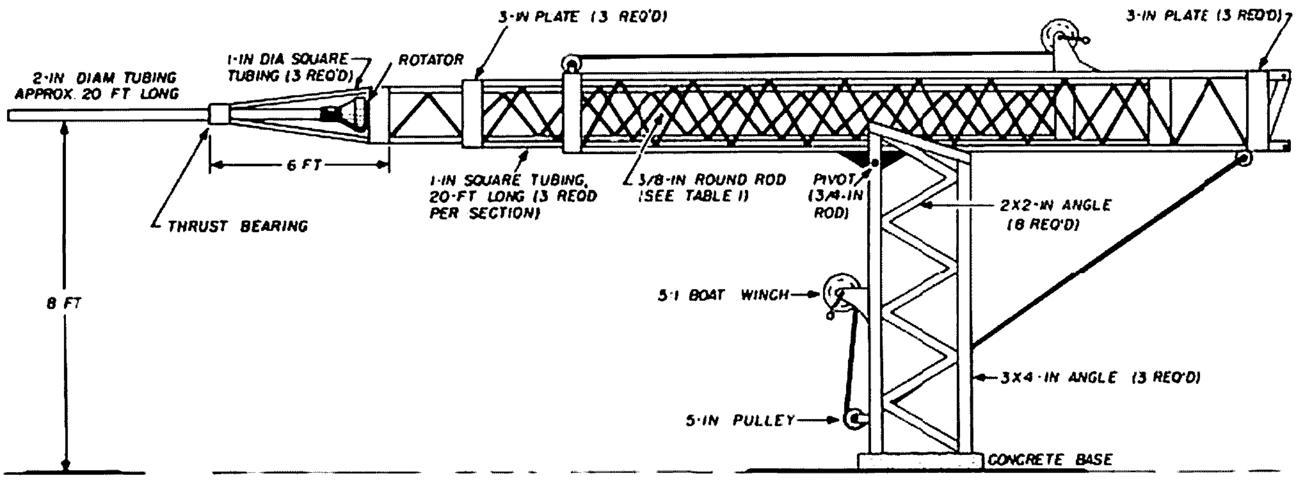
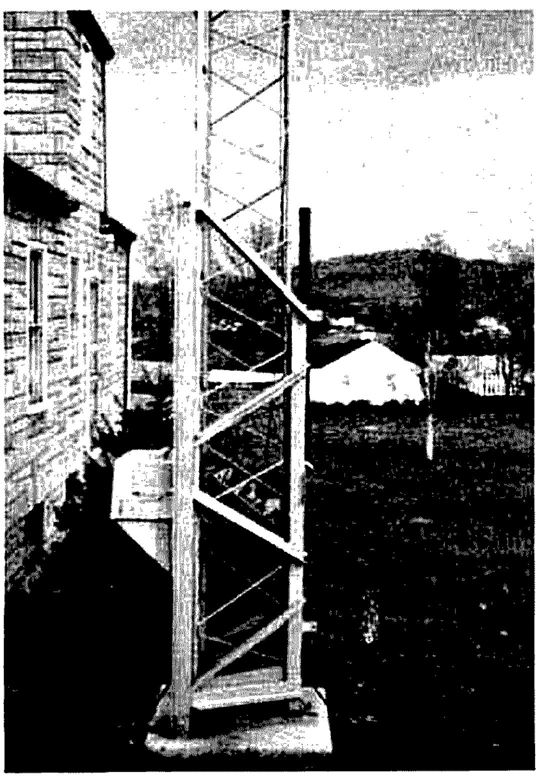


fig. 1. The 70-foot tower in the tilt-over configuration. The tubing extension at too is adequate for a 6-meter yagi antenna; other antennas will require extensions using different tubing sizes and lengths.



Clean lines and true verticality result from careful jig measurement and prealignment before welding and placement in concrete base.

templates

My reply to WA3AHM's challenge about building a triangular tower was, "You can't keep the triangular elements

table 1. List of materials for the homebrew tilt-over tower. All material is soft iron.

quantity	description	size
1	round	3/4" x 25"
2	angle	3" x 4" x 8' 6"
1	angle	3" x 4" x 6' 8"
8	angle	2" x 2" x 27"
66	round	3/8" x 23"
66	round	3/8" x 22"
66	round	3/8" x 21"
10	square tube	1" x 20'
18	plate	1/16" or 1/8" thick x 3" wide; approx. 19" long

*An estimate for the complete welding job, based on precut and accurately dimensioned parts, is about $50.00 for labor and welding materials. With some energetic bargaining, this cost could be reduced. editor.

aligned during construction." Then I came up with the idea of making three jigs, or templates, from one 4 x 8 sheet of 5/8-inch-thick plywood. The detail of the jigs is shown in fig. 2.

After laying out one piece of wood in the hexagonal shape shown, I made two more and nailed all three pieces together. Three circles were inscribed on the top piece, then each circle was divided into six equal parts using a large pair of dividers.

Next, one-inch squares were laid off at each corner of the three triangles. These were drilled with a ¾-inch bit and shaped into square holes with a saber saw. (These holes will accept the main longitudinal members, which are 1-inch-square lengths of tubing. See table 1 and fig. 1.) The 1-inch-square holes should allow the tubing to slide through the wood easily, but with a slight amount of drag. It might be necessary to stagger the holes slightly in the center piece to provide solid support for the tubing in the jigs.

setup

The jigs were separated and numbered 1, 2, 3 in relation to the way they had been nailed together. The 1-inch tubing was then placed through the holes in jig 2 and allowed to extend about six feet beyond the end of the jig. Jig 1 was placed just over the end of the six-foot protrusion of the tubing, and jig 3 was placed over the far end of the tubing. It is important that the number of all jigs face in the same direction. Otherwise accumulative alignment errors will creep in, which will affect the verticality of the final structure. (You don't want the neighbors to think you're a CBer!)

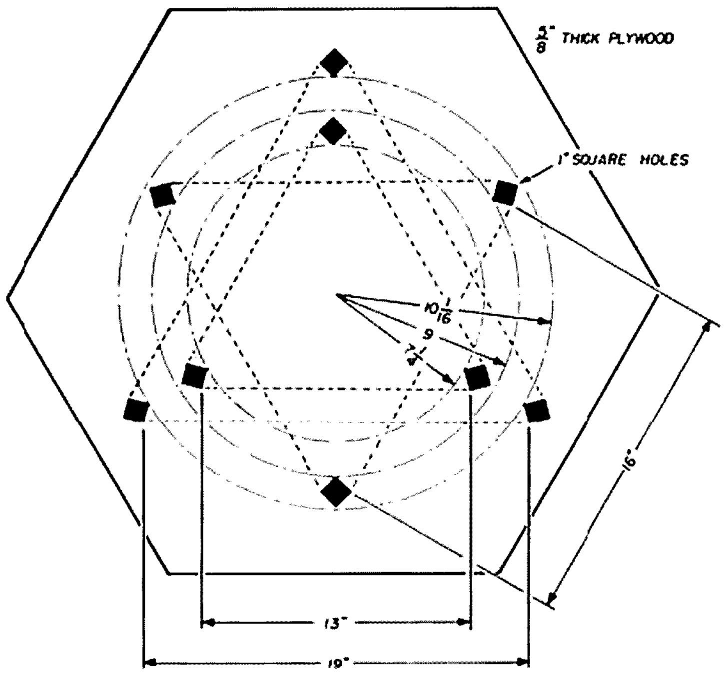
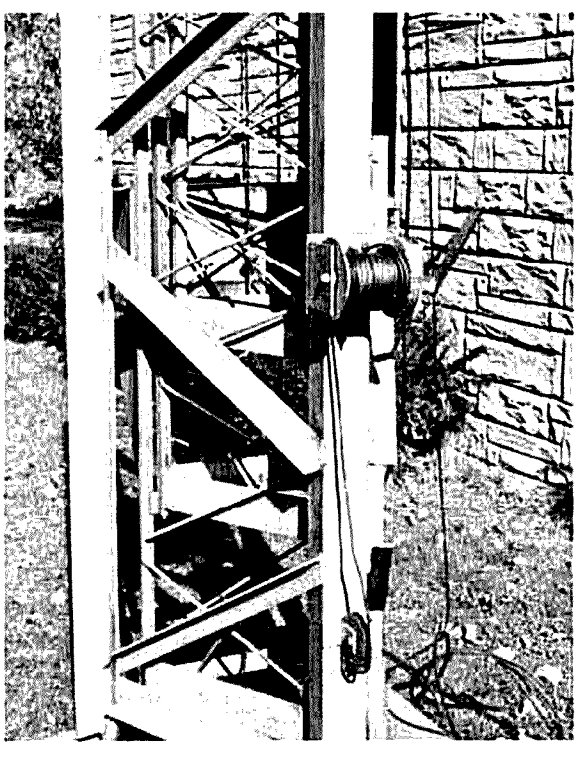


fig. 2. Geometry of templates used for positioning tower elements during welding. Jigs are made from a single sheet of 4 x 8, 5/8-inch plywood.

Details of the tilt-over stub and boat-winch.



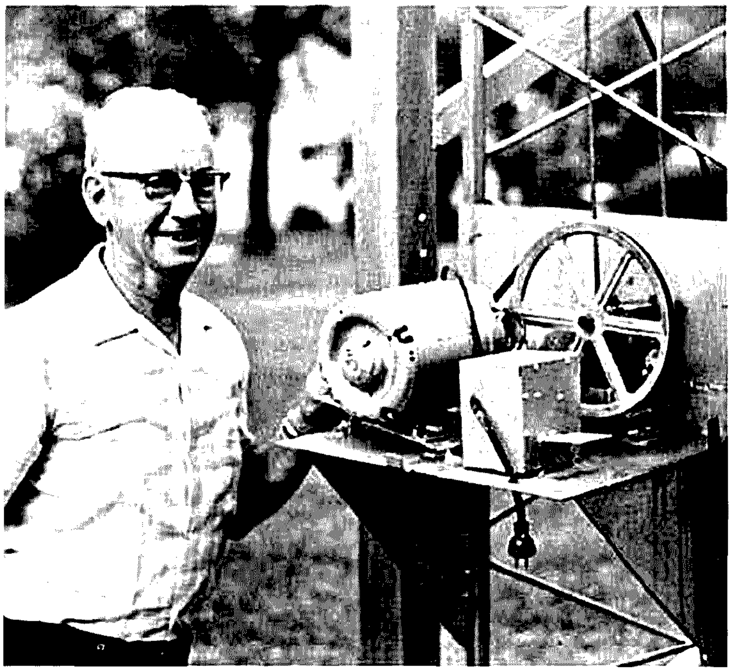
Choose a level spot on which to set the jigs and tubing. You'll need some wooden wedges to level the jigs. Place a string over the tops of jigs 1 and 3, and adjust jig 2 for level if necessary. Then run the string along the corner of the jigs to make sure there is no side twist in the assembly.

welding

The 3-inch plates were welded around the triangle at jig 1 first. Then the 3/8-inch round rods were placed zigzag fashion up the side of the triangle and tack welded. This procedure was repeated on all three sides of the tower section, up

to movable jig 2.

A problem in welding a structure of this type is positioning the diagonal truss members and keeping them in place while welding. I used a magnet salvaged from the yoke of an old TV set to hold the pieces in place during the tack welding.



Electromechanical elements for raising and lowering the tower. A 1/3-HP motor and 40:1 gear train are used.

Each diagonal truss was bent slightly (approximately ½-inch radius) before welding. This enhances the appearance of the structure.

The next step, after tack welding the diagonal members, was to move jig 2 approximately six feet and recheck for level and twist. The tack welding was continued until jig 2 was against jig 3. Then jig 3 was removed, and jig 2 was left in place. The 3-inch plate was welded around this end of the triangle. Next, both jigs were removed, and the 3/8-inch round rods were welded permanently. This completed one section of the tower; the other two sections were constructed similarly. Three-inch-long pieces of ¾-inch angle were then welded at the corners of the sections to act as guides when the sections are raised and lowered.

tilt-over stub

This part of the assembly was built from 3 x 4-inch angles (longitudinal pieces) and 2 x 2-inch angle trusses. The tilt-over stub is triangular and mounted in a concrete base. Use extra care here to obtain an absolutely vertical structure. Make several measurements, then recheck each before the concrete is poured. This work is extremely important if the tower is to be vertically true.

A length of ¾-inch round rod completes the assembly as a pivot piece for the tilt-over stub.

accessories

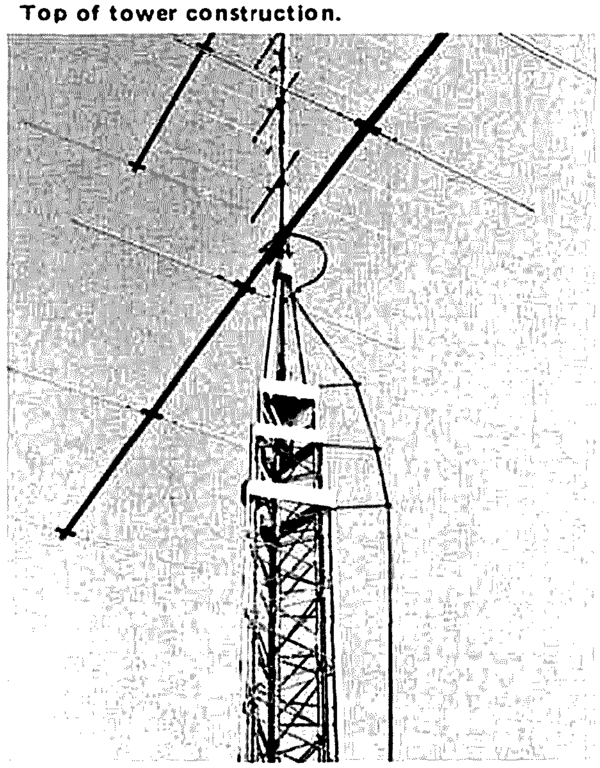
The lifting and lowering mechanism consists of a 1/3-HP electric motor and a 40:1 worm-and-gear arrangement. The tower is tilted by a 5:1-ratio boat winch and pulley assembly. The cables for the entire tower are 1/4-inch galvanized stock.

acknowledgment

Several amateurs in this area have used these jigs to build similar towers. I'd like to thank all who have contributed information for this article.

ham radio

Top of tower construction.



the double bi-square array

The ressurection of an old classic that provides high performance on the high-frequency bands

Bernie Ontiveros, W6FFF, 118 South Alisos Street, Santa Barbara, California 93103

The bi-square antenna was originally described in the April, 1938 issue of *RADIO* magazine by Woody Smith, then W6BCX and editor of the magazine. I was working for the magazine at the time and thought I would build one some day. Now I have, more than 31 years later!

The original configuration with one point downward is necessary for the single pole mounting, but the same antenna is shown in W6SAI's "Quad Handbook," with one side parallel to the earth as used in the conventional quad arrangement. This is called the XQ antenna, meaning Expanded Quad.

The bi-square array, a derivation of the Lazy-H, has a broadside gain of about 5 dB with a bi-directional pattern. It is twice the size of the conventional quad element, as each side is a half-wave long. This means that it is hardly practical to rotate except on 10 meters. On that band some have been built with a reflector or director with a resultant unidirectional pattern and gain of around 9 dB. KV4AD has had a big signal on 10 meters with a variation of this rotatable arrangement.

A 10- and 15-meter pair of bi-square arrays may be mounted concentrically; with another similar pair mounted at right angles. Instant switching of direction and frequency may be accomplished at the operating position as shown in **fig. 2.**

construction

The dimensions of the bi-square array are not too critical as the simple matching method shown here resonates the antenna sufficiently when tuned for minimum swr. Each side of the 10-meter array is about 16½ feet long, and each side of the 15-meter array is about 22½ feet long. The open-ended stubs are 8½ feet long for 10 and 11½ feet long for 15.

This is not a high-powered antenna as shown here. The plastic insulators are unavoidably at points of high rf voltage and RG-58/U coaxial feed line is used for convenience. This is satisfactory in my case since the longest feed line is less than 20 feet, and the maximum power is that attained by a 500-watt PEP (input) trans-

ceiver. High powered operation will call for better insulation and RG-8/U feedline.

The whole system is hung on a lightweight wood pole consisting of a pair of 12-foot 2X4s spaced with blocks of 2X4 as a fixed base unit. A 2X3 is butt-spliced to a 2X2 top section. The bottom of the 2X3 is inserted between the 2X4s and fastened with two bolts. With one bolt in to act as a hinge, the whole upper part is pushed up, and the second bolt put in to secure the mounting.

Four nylon lines, about 2/3 up the pole will suffice for guys during installation. After the arrays are in place they will serve as the final guying.

It is a good idea to make a scale drawing of the antenna to locate the placement of pulleys for symmetrical rigging. Those for the 15-meter antennas at the top should be offset a few inches vertically and on adjacent sides of the pole to avoid entanglement. The same applies to the pulleys for the 10-meter antennas at the plotted distance down the pole.

The pulleys are small aluminum types with nylon rollers, with eyebolts you put on and enough bolt length to go through the pole. The nylon line is a utility type that comes in 100-foot hanks and has a diameter of slightly less than 3/16 inch.

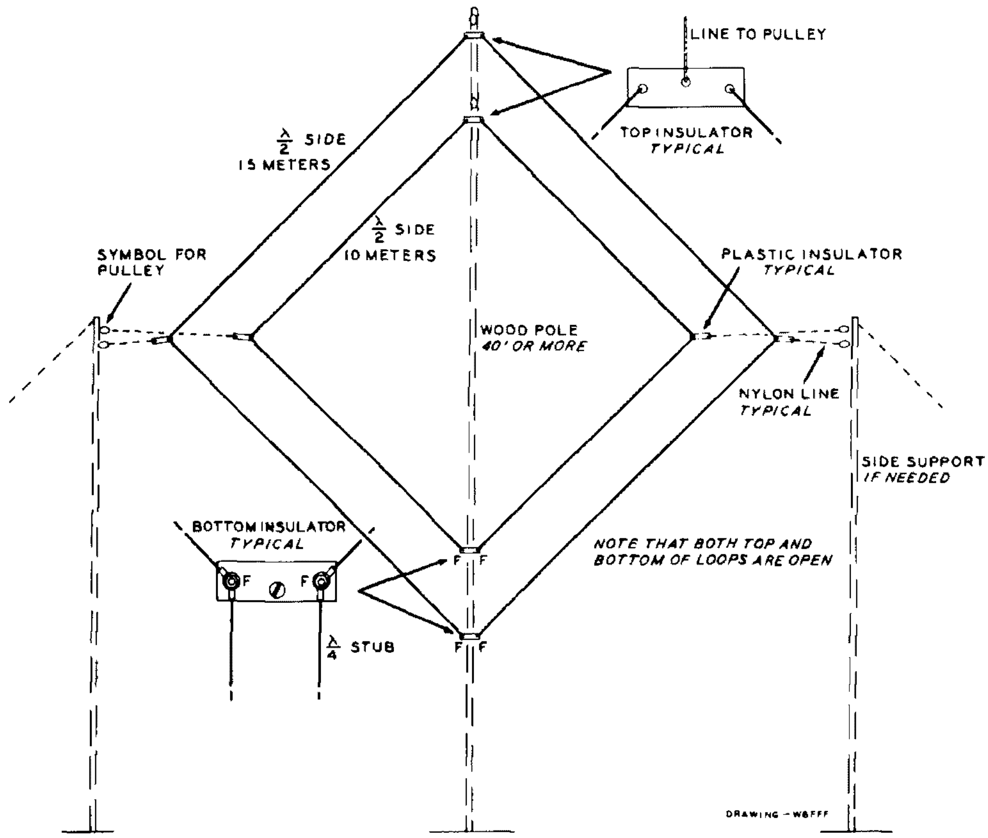


fig. 1. Bi-square array for 10 and 15 meters for one bi-direction. For doubled directional coverage a similar array is hung on the same pole at right angles.

The top 15-meter antenna goes up first, the sides are pulled out, the bottom placed, and the whole thing pulled snug. The others follow in the logical order.

(The drawings in the antenna books always show perfect squares with side guys going out to an invisible fence or ground stake. However, with my 50 foot wide lot the best I could achieve with fences was a skinny diamond. Some 20-foot sections of tv mast squared up the arrays.)

Now the stubs may be installed, straight out and tight, trying to keep the 4 stubs as close to 90 degrees from each other as possible. Install the coaxial feedlines and tune up.

tuning

Tuning up is simplicity itself. Slide the coaxial line connector to the point of minimum swr near the center of the band and that's it. This will hold pretty well over the whole band with acceptable swr since this is a fairly low-Q antenna.

Further sophistication may appeal to some, such as resonating the whole antenna with a shorting bar and then finding the spot for attachment of the feedline, perhaps with some type of balun. After several months use I put simple bazooka quarter-wave sleeves on two of the antennas, but as no difference was noted, good or bad, I just left them on.

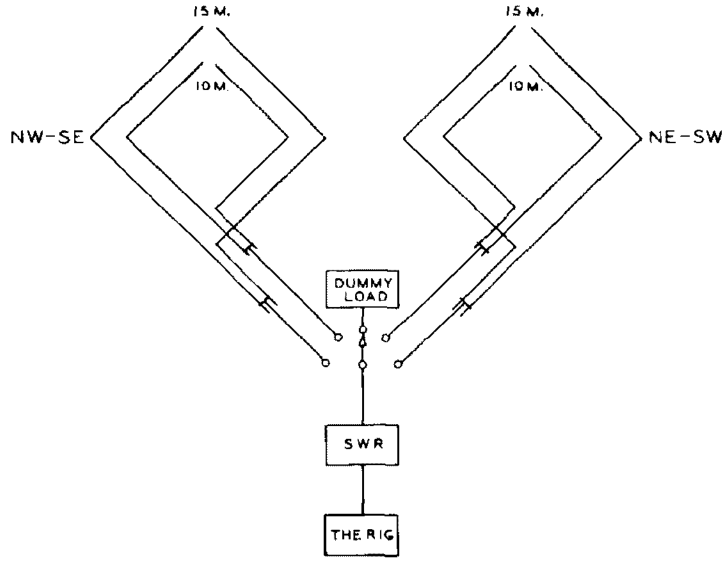


fig. 2. The full array. This is four separate antennas; any single one may be used since there is no dependence on any of the others.

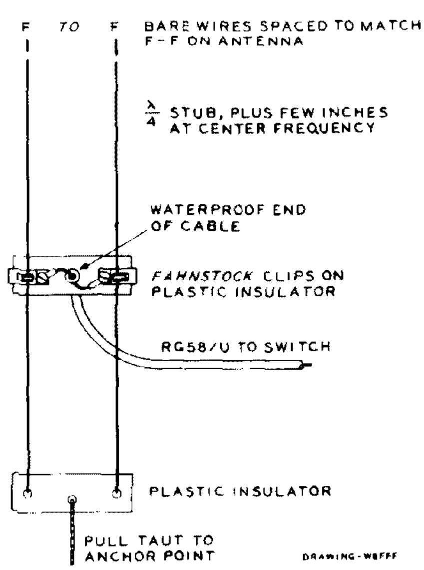


fig. 3. The movable feed point. After positioning the clips may be soldered to the wire and the whole thing covered with a waterproof cover.

performance

After a year's use a pretty good evaluation of the antenna is possible. The instant switching of directions is interesting as it often shows a 5 or 6 S-unit difference between the antennas, both locally and distant. The 10-meter antennas show more discrimination than the 15, probably due to better electrical spacing above ground. The exact center between beam paths doesn't seem too lively, but that may be due to areas of lesser ham activity, as I have received some good reports along those paths from Antarctica, South Africa and VK9.

My particular setup of SW-NE, through New Zealand and Spain, and NW-SE, through Japan and Argentina, seems to favor those areas.

A good long-term check on this antenna has been possible through almost daily contacts over the past year with ZL3LE on 10 meters, and with ZL3KA on 15 meters through all sorts of conditions. Both Bill and Jack agree that I am competitive with any of the W6s except for certain well known big antenna operators, and they of course are also running 2 kW PEP which helps a bit.

ham radio

uhf transmission lines

For maximum performance on 432 MHz and above you must select low-loss transmission lines — here are some of the types most suitable

Dolph Vilardi, WA2VTR, 14 Oakwood Terrace, Spring Valley, New York 10977

Most discussions of transmission line for amateur radio seem to describe the various types available, leaving the ultimate choice to individual decision. While this may be a satisfactory approach for the high-frequency bands, and possibly vhf, at uhf (432 MHz and above) the choice is narrowed down to a very few types. Nominal losses inherent in the chosen transmission line plus impedance mismatch bring total losses to a point which is completely unacceptable.

Assuming a reasonably good test setup, let's consider the example of a 1296-MHz antenna with an unknown mismatch (ordinarily considered acceptable at high frequencies). The antenna is fed with 80 feet of *good quality* RG-8/U line. When 60 watts are fed into the line at the transmitter, a reflected power of 1 watt is indicated at the transmitter (which would appear to be a fairly acceptable vswr).

Now let's see what we actually have. Eighty feet of RG-8/U has 8-dB line loss at 1296 MHz. With this loss 9.6 watts of power arrives at the antenna. Remember that we *measured* 1 watt reflected. This means that the actual power reflected from the antenna was 6 watts. Therefore, the actual power radiated by the antenna is the difference between the power that arrived and the power that is reflected. Since 9.6 watts arrived, with 6 watts reflected, the effective radiated power is 3.6 watts. Clearly this is unacceptable.

The solution to this problem lies in as nearly perfect matching of the transmission line to the antenna as possible and the use of low loss line.

Table 1 shows some of the lines that may be used at uhf. If in the previous example we had matched the transmission line to the antenna, used a line which had a nominal loss of 3 dB and measured one-half watt of reflected power we could have the following: 60 watts at the transmitter, 3-dB line loss (30 watts at the antenna) and ½ watt reflected power (1 watt is reflected back at the antenna).

Thirty watts minus 1 watt equals 29 watts effective radiated power, an eight-fold increase compared to the previous example. All with the same transmitter. The cost of the transmission line would be far less than the cost of quadrupling transmitter power. Moreover, the equivalent gain would be obtained at the receiver since an antenna system is reciprocal on receive and transmit.

From the standpoint of nominal loss the poorest line that should be used at these frequencies is equivalent to *good quality* RG-17/U. Remember that poorly made line may have impedance bumps due to variations in concentricity, dialectric quality and braid characteristics. Quality is important.

A solid center conductor is mandatory. Of course, coaxial fittings should be *those made for the line.*

Manufacturer's specifications regarding minimum bending radii should be carefully followed since distortion of the

table 1. Losses for coaxial transmission lines suitable for amateur use. RG-8/U is included to show how poor it is in comparison with other types.

type	O. D.	shield	jacket*	attenuation per 100 feet (dB) 432 MHz	1296 MHz	power loss (1296)
RG-8A/U	0.405	single (braid)	PVC-2	7	12	93.7%
RG-9/U (RG-214/U)	0.420	2 (braid)	PVC-2	7	12	93.7%
RG-14/U	0.545	2 (braid)	PVC-2	3.8	7	80%
RG-17A/U (RG-218/U)	0.870	1 (braid)	PVC-2	2.4	4.5	67%
RG-19/U	1.120	1 (braid)	PVC-2	1.8	3.6	55%
RG-231/U	0.500	alum tube	none	2.1	3.5	55%
RG-331/U	0.500	alum tube	PE	2.1	3.5	54%
RG-360/U	0.750	alum tube	PE	1.7	3.0	50%
RG-332/U	0.875	alum tube	none	1.4	2.5	42%
RG-333/U	0.875	alum tube	PE	1.4	2.5	42%

*PVC-2 indicates non-contaminating polyvinyl chloride; PVC-1, contaminating polyvinyl chloride is found on regular RG-8/U and should not be used. PE is polyethylene.

transmission line can cause serious impedance disturbances at 1296 MHz. Specifications should also be adhered to regarding burial of lines. It should be realized that plastic is permeable to water vapor at varying rates; pinholes in the plastic covering are nearly unavoidable in manufacture.

Some of the gas-filled lines have excellent nominal loss characteristics but unless some means of pressurizing is available stay away from them since moisture inevitably gets in and destroys the low-loss characteristics. Unless pressurization is available stick with foam-filled lines.

Twinlead is often recommended for uhf because of its *nominal* low-loss characteristics, but in actual use it is quite undesirable since it requires great care in dressing, insulation from surrounding structures and the strict avoidance of sharp bends. In addition, line radiation is a problem, as is impedance matching. When it rains the line becomes practically useless. My advice is to forget it.

It should be mentioned that the short piece of flexible line around the rotator often represents a fairly large proportion of the total line loss in a good system. Use the best you can get such as double-braided RG-9/U or better yet, RG-14/U or Times FM8. Remember that 7 feet of RG-8/U around the rotator represents 0.7 dB or nearly 10% power loss.

Of course, waveguide has excellent characteristics at these frequencies and makes excellent low-loss line. However, the same problems of pressurization occur as in air-supported coax although not to as great an extent. Also, problems at corners and bends and transformations to coaxial cable (for use with a rotator) present certain design problems. High cost is another factor which rules it out.

In conclusion some mention should be made of the type of coaxial fittings you use. It goes without saying that they should be constant-impedance types. This automatically rules out the commonly used series-uhf fittings such as the PL-259. I strongly suggest that you standardize your uhf system to N-type fittings.

N-type fittings are constant-impedance types and have the added advantage of being fairly weatherproof. In addition, the N-type connector's low-cost availibility on the surplus market makes it even more attractive. It is made in a wide variety of fittings so it can be used with almost any transmission line which you choose. Its characteristics at 1296 MHz are excellent, and it can carry a great deal of rf power without breaking down.

It should be obvious from the above discussion and **table 1** that under all conditions the transmission line should be as short as possible.

ham radio

low-cost compact antennas for 20 meters

A selection of simple antennas based on plastic-pipe booms and element supports for the antenna experimenter

John McFarland, W4BOS, 12 Sandra Drive, Port Richey, Florida 33568

All compact antennas are compromises, whether they are dipoles, verticals or beams. For low swr they must be operated over rather narrow frequency ranges, and efficiency is not as great as with a full-sized antenna. However, by using elements at least 1/8-wave long, inefficiencies may be kept to a minimum.

In the *tripole* antenna I have tried to overcome some of the disadvantages of miniature antennas by adding a third element as shown in **fig. 1.** Although this antenna looks like a ground plane at first glance, it is not; it consists of an inverted-vee dipole with a vertical element connected to one side. Performance of this arrangement has been excellent, and I have received good signal reports on both 20 and 40 meters.

The vertical support may be made from varnished bamboo or ½-inch plastic pipe. If you use plastic pipe, use plastic fittings to couple the sections of pipe together. The vertical element for the 40-meter version (originally described in *Florida Skip*) is shown in **fig. 1.** The vertical element for the 20-meter tripole consists of a 16-foot, 6-inch section of number-14 wire taped to a 17-foot bamboo (or plastic) pole; the inverted-vee elements of the 20-meter version are 16½-feet long.

The vertical element and one of the inverted-vee elements are connected to the center of the coaxial feedline; the other inverted-vee element is connected to the outer braid of the coax. To tune the antenna, use a grid-dip meter to indicate resonance. In the 40-meter tripole the windings of loading inductance L2 are expanded to increase frequency, and compressed to lower resonant frequency.

tripole beam

Two compact tripole antennas may be combined into a beam as shown in **fig. 2.** Although this particular antenna was designed for 20 meters, similar designs could be used on the other bands. The compactness is particularly useful for space-cramped amateurs who want to operate on 160, 80 and 40 meters.

In the 20-meter tripole beam, the boom is a 1-inch square aluminum pipe, 101-inches long. Each of the tripole elements are 8-feet, 8-inches long. Construction is shown in **fig. 2.** Each of the elements is center loaded with coils wound on a 5-inch section of plastic PVC pipe 3/4-inch in diameter; fill the form with 4 turns number-18 plastic-insulated wire, spaced 1/16-inch. The two 48-inch aluminum tubing elements are held together with 12-inch lengths of ½-inch PVC pipe.

Each aluminum element is pushed 2 inches over the end of the center insulator, leaving 46 inches of active element. The plastic coil form slides over the 1-inch center insulator and is connected to the aluminum elements with short pigtails. (Note that all coils are wound in the same direction.) This movable coil form is used to tune the antenna to resonance by moving it to a different position on the center insulator.

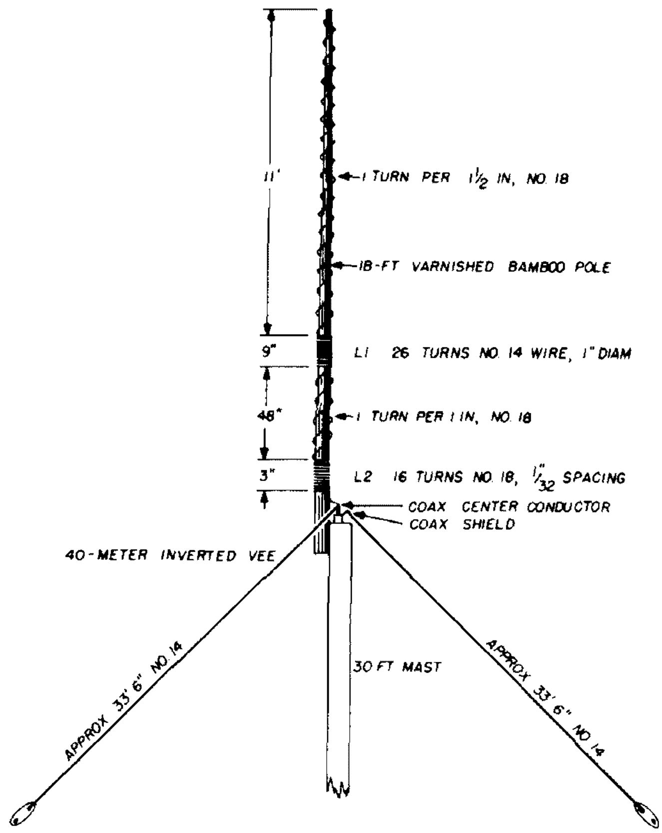


fig. 1. Tripole antenna for 40 meters. Grid-dip the vertical element to 7250 kHz. Adjust resonance with L2.

Once each element has been resonated, the movable coil form is cemented to the center insulator. This is a rather high-Q antenna so one frequency setting will permit operation over 100 kHz of the 20-meter band for swr less than 1.5:1.

Each of the six tripole elements is held to the boom with plastic tee fittings, available at your local hardware store. When all the elements are in place on the boom, connect the base of the reflector elements together with a 4-turn coil of number-14 wire, 1½-inch in diameter, 2-inches long. The radiator is fed with a ¼-wave matching transformer made from two sections of RG-58/U; the 50-ohm coaxial line from your transmitter is connected to the bottom of the matching transformer as shown in fig. 2. One center conductor of the matching section is connected to one of the horizontal elements; the other center conductor is attached to the vertical element and the remaining horizontal element.

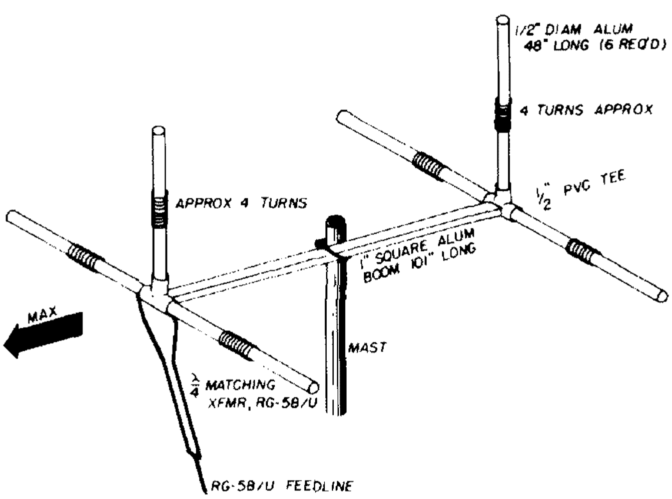


fig. 2. Tripole beam uses plastic-pipe and aluminum elements. All plastic elements shown in fig. 3 can also be used.

plastic-pipe tripole beam

The basic tripole system shown in **fig. 2** is easily, and less expensively, made with PVC-pipe supports and wire elements. For this simplified construction, use an long, may also be used for the boom. Each of the tripole elements is mounted on a plastic tee-fitting which is mounted to the boom. The three elements of the reflector are connected together with a

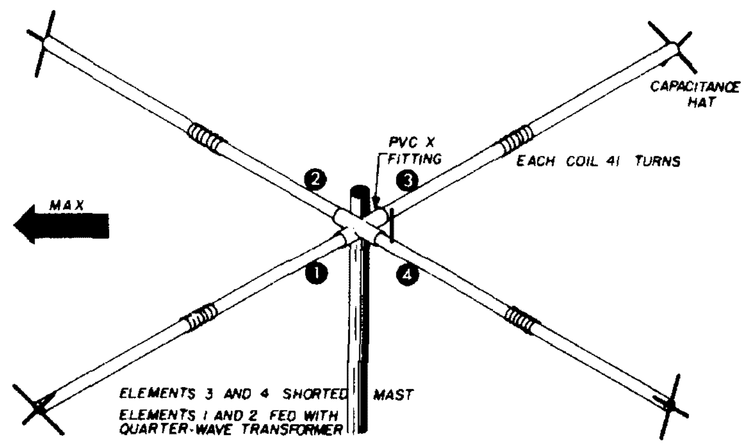


fig. 4. X-bar antenna uses four elements of the type shown in fig. 3. This antenna may be used horizontally, as here, or vertically.

8-foot, 8-inch length of plastic pipe for each element. The loading coil is wound on the outside of the tubing, and after tuning, is cemented in place. Drill a small hole in the pipe on each side of the coil, and connect the coil to 4-foot sections of wire inside the pipe as shown in **fig. 3.** Plug the ends of the pipe with wooden dowels to keep moisture out.

Each of the loading coils consists of 30 turns number-18 plastic-insulated wire on a 1-inch diameter form with a winding

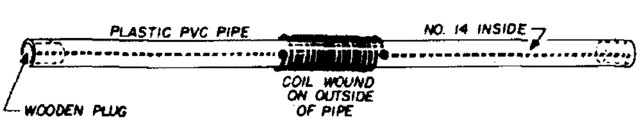


fig. 3. Plastic-pipe element consists of center loading coil and wire elements inside the pipe.

length of 7 inches. This coil, with a capacitance hat made from two 12-inch lengths of number-14 wire, provides resonance on 20 meters. Tuning range of this all-plastic antenna is on the same order as the more conventional design shown in **fig. 2.**

A section of plastic pipe, 101-inches

coil and tuned 5% lower than the radiator. The radiator is fed in the same way as shown in **fig. 2.** The tripole elements may also be mounted to the boom with plastic X fittings so the elements are symmetrically spaced 270° apart.

x-bar beam

The X-bar beam antenna is an extension of the tripole beam that uses four compact elements instead of three. Element construction is the same as that shown in **fig. 3.** The radiator and reflector each consist of four elements, attached to the 101-inch plastic boom with plastic cross fittings. The X-bar antenna is fed with a matching section as shown in **fig. 2.** Performance is slightly improved over the three-element version.

x-bar compact antenna

The X-bar compact antenna shown in **fig. 4** was developed as a result of experience with the X-bar beam. This antenna consists of four PVC pipe elements, and may be used either horizontally or vertically. I mounted one version of this antenna on a short tilt-over mast extension which could be lowered

for vertical polarization, and put up (like an umbrella) for horizontal operation. Under some skip conditions I found that switching from vertical to horizontal increased signal strength. However, the swr

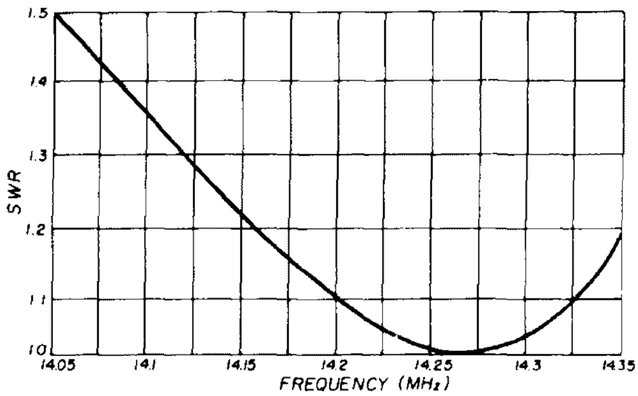


fig. 5. Swr performance of an X-bar antenna using plastic and aluminum elements.

changes from horizontal to vertical, with a slightly higher resonant frequency in the vertical position.

Each of the elements in the X-bar antenna is the same as those shown in **fig. 3.** For vertical operation the coils should be 30 turns; for horizontal operation, the coils should be 26 turns, number-18 plastic-insulated wire. With a horizontal mounting the quarter-wave matching transformer was made with 75-ohm RG-59/U; swr was nearly flat for 50-kHz each side of resonance, with an swr of 1.3:1 for a total frequency range of 150 kHz.

The X-bar antenna can also be built with the plastic and aluminum elements shown in **fig. 2.** The swr performance of the aluminum X-bar antenna with capacitance hats is shown in **fig. 5.** These measurements were made with the antenna 9-feet off the ground. For tuning it is much easier to prune the capacitance hats than to adjust the loading coils. The capacitance hats on elements 1 and 2 are 12-inches long; the element hats on elements 3 and 4 are 6-inches long.

For better swr performance two elements of the vertical X-bar antenna can be combined with full-size quarter-wave wire elements as shown in **fig. 6.** This antenna is much more directional than a dipole, and provides much broader frequency coverage. The swr is less than 1.5:1 for the entire 20-meter band. The matching transformer used with this antenna is the same as that shown in **fig. 1.**

antenna tinker kit

After building the various antennas described so far it occurred to me that these same elements could be combined in any number of ways to provide useful antennas. The elements can be built in several ways, including the aluminum and plastic-pipe version in **fig. 2** and the lower cost all-plastic-pipe version in **fig. 3.** There are several other ways of making elements, including wrapping the ends of the plastic pipe with aluminum foil and

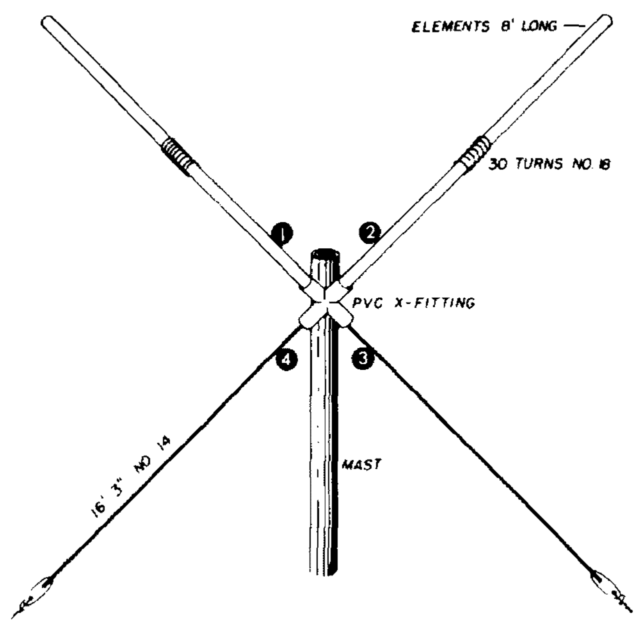


fig. 6. Improved performance is provided by two compact elements with two full-sized quarter-wave wire elements. Elements 1 and 2 connected to 1/4-wave transformer; elements 3 and 4 connected to shield.

doing away with the internal wire elements. To connect the loading coils to the aluminum foil, arrange the pigtails in the form of a sine wave and place them on top of the first layer of foil; put another layer of foil over the pigtails and tape it in place.

Another method of making elements is shown in fig. 7. In this element holes are drilled along the length of the plastic pipe and aluminum ground wire is woven in and out of the pipe as shown in the drawing. Each section of wire is 7-feet long. With this arrangement the loading coil is smaller; about ten turns around the ½-inch pipe is about right for 20 meters.

Although these elements are shown without capacitance hats, hats can be used. I have used the wire hats shown in fig. 2. as well as the loops shown in fig. 8. The loops in fig. 8 are made from a 22-inch piece of number 16 wire. The antenna element is tuned by changing the length of the loop. With the capacitance loop, the center loading coil consists of 42 turns number-18 plastic-insulated wire for resonance on 20 meters.

fig. 7. Another method of making low-cost antenna elements.

These compact elements may be arranged in many different configurations, including horizontal and vertical dipoles, inverted vees, upside-down inverted vees, X-bars, X-bar beams, tripole beams, conventional beams, ground planes, and various phased arrangements. Because of their low cost and ease of construction, this type of element is ideal for the antenna experimenter who wants to try different types of antennas.

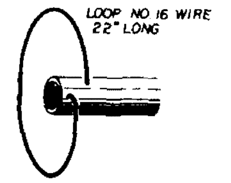
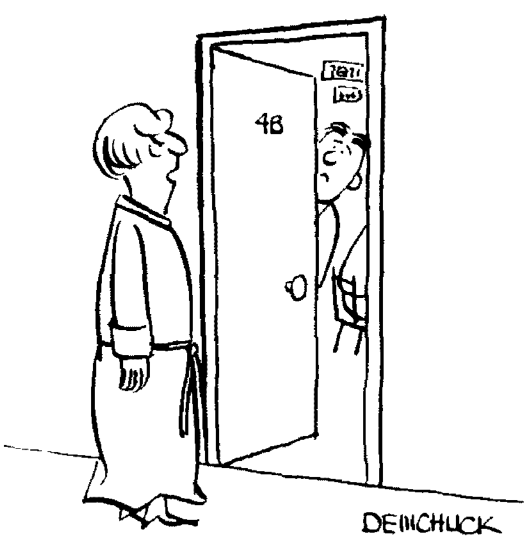


fig. 8. Capacitance hat made with a loop of wire. Resonance is adjusted by changing the length of the loop.



"Did a man come up here to complain about radio interference two hours ago?"

tuning

To tune each of the compact elements you need a grid-dip meter and a full-sized half-wave dipole about 5-feet off the ground. First, grid-dip and dipole and adjust it for the frequency at which you want to operate. Mark the point on the grid-dipper dial. For best accuracy use a communications receiver to check the calibration of the grid-dip meter. Use a 2-turn link to couple the grid-dip meter to the dipole.

Now connect the compact element to one side of the dipole and prune the capacitance hat (or adjust the loading coil) until the system resonates at the same frequency as the dipole did by itself. Resonate the rest of the elements one at a time with the same system.

To tune the coils you can temporarily connect a 50-pF trimmer capacitor across the loading coil of an element that has been previously resonated with the wire dipole. Without disturbing the trimmer or the grid-dip meter, connect the trimmer across the loading coils of the other elements and adjust the coils for the same resonant point on the grid dipper. This way you know the coils in all the elements have the same inductance.

For best results use white PVC pipe instead of black. The white reflects the rays of the sun and holds up much better in the weather.

ham radio

type-F coaxial-cable fittings

The case for 75-ohm transmission line in amateur applications

Serge Ticknor, K2MDO

The type-F coaxial connector has been the standard fitting of the Community Antenna Television (CATV) industry for several years. Exactly why hams haven't adopted these low-cost, high-performance coax fittings is not clear. I personally believe it's because of a general lack of awareness of their adaptability to amateur use.

The F fittings are available for 75-ohm cable only. In many cases, 75-ohm cable is preferable to 50-ohm cable for at least three reasons:

1. The impedance is more adaptable to the popular ham antennas (an exact 4:1 impedance ratio for 300-ohm folded dipoles using a balun feed, or 1:1 for a center-fed dipole or coaxial vertical).

2. Less attenuation is offered by 75-ohm cable than in an equivalent size and length of 50-ohm cable (an important consideration at the higher amateur frequencies).

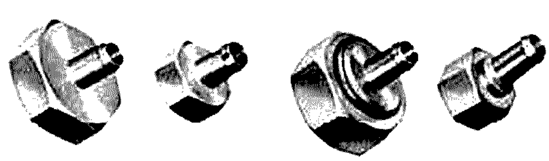
3. 75-ohm cable is widely used in CATV and many home TV installations, making it economical and readily available.

advantages of F fittings

Advantages of F fittings are derived from their very reason for being developed. Cable television required a fitting with good rf characteristics throughout the uhf range, ease of installation, and economy. Before the development of the F fitting, connectors in general use were the automobile type, uhf, type N, and type BNC; each with its limitations.

The auto-type connector has obvious high-frequency and mechanical limitations. The uhf connector, still used by many hams, is rugged but bulky. Great care must be taken during assembly, particularly when soldering the coax braid to the connector shell. The frequency limitations of the uhf connector are well known despite its name. Many years ago, when it was first developed, the term "uhf" meant the region around 200 MHz.

Type-N and BNC connectors have fairly good high-frequency characteristics and are waterproof, but cost is high. These cable fittings consist of several intricate parts that must be carefully assembled to the cable. The parts include a center pin, shell, sealing washer, ferrule, braid compression ferrule, and clamp nut. Assembly or salvage of these fittings is difficult and must be done with great care.



Blonder-Tongue coax connectors BTU-590, BTF-590, BTU-591 and BTF 591 (from left to right).

the F connector

A typical type-F cable fitting, such as the F-59A, consists merely of a short mandrel permanently attached to a swivel nut, with an integral cable-clamping ferrule attached. The cable is stripped and pressed into the fitting, and the ferrule is compressed onto the cable. The center conductor becomes the center matching pin; and the connector, instead of the braid, becomes the outer conductor.

With this design there is negligible impedance discontinuity, even through 1000 MHz. This is because near-identical physical relationships exist between the cable outer braid and inner conductor, and a second dielectric hasn't been introduced to affect the impedance of the assembly. Internally, the connector is merely a physical continuation of the cable; so a near-perfect transition exists.

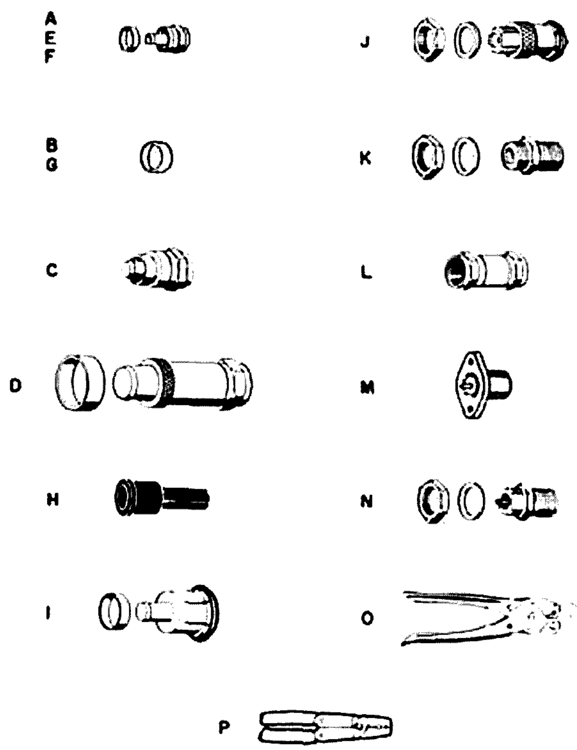


fig. 1. Available type-F connectors and accessories (items are identified in table 1).

A variety of fittings is available (fig. 1 and table 1). Chassis jacks are also available, as are quick-disconnects and splices. All are available for RG-6/U, RG-11/U, and RG-59/U cable. In addition manufacturers supply switches, variable attenu-

table 1. Type-F fittings and accessories

sketch (fig. 1)	description	Jerrold	Blonder Tongue	Channel Master	Winegard	JFD
	Type-F Cable Fittings for RG-59/U					
A	Swivel type, male; uses center conductor of cable for center pin; mates with any F-type jack	F-59	BTF-591	—	F-59	C-7106
B	Cable crimping ferrule for above connectors	1051	HR-59	—	F-59	C-7113
C	Similar to A with built-on ferrule	F-59A	—	7194	—	—
D	Waterproof male connector	—	BTU-592	—	—	—
	Type-F Cable Fittings for RG-11/U					
E	Swivel type, male; uses center conductor of cable for center pin	—	BTF-110	—	F-11	C-7109
F	Swivel type, male; has its own center pin	AF-101	—	7196	—	—
G	Cable crimping ferrule for above connectors	1059	HR-11	—	F-11	C-7114
	Weather Boot					
H	Waterproof weather boot for all RG-59/U and RG-11/U fittings	WB-56	—	7197	—	—
	Type-F Push-on Quick Disconnect Cable Fittings					
I	Connector uses center conductor of cable for center pin; constructed for direct insertion of cable	—	BTF-59P	—	—	C-7116
J	Connector uses its own center pin; requires additional attachment to F-type fitting on cable	F-91	QDP	—	FP-59	—
	Adapters					
K	Double female coupling adapter; couples two male F-type connectors together	F-81A	GF-81	7195	F81-M	C-7108
L	Double male coupling adapter; couples two female F-type connectors together	F-71	AD-3	—	—	C-7110
	Chassis Panel Fittings					
M	Female chassis connector; has two rivet or screw holes for mounting	—	BTF-100	—	—	—
N	Female chassis connector; 3/8 in. dia. mounting hole with nut	F-61A	—	7193	—	C-7105
	Crimping Tools					
O	Heavy duty, continuous parallel jaw chuck for RG-59/U and RG-11/U	PL-602	—	7188	—	—
P	Economy type, scissors-jaw pliers	PL-659	CR-2	—	CR-1	—

ators, and other devices usable by hams — all adaptable to type-F fittings.

weatherproofing

A weatherproof connection is only required outdoors, so the extra cost of a weatherproof fitting such as the type-N isn't justified for indoor use. Some manufacturers supply a rubber or plastic boot for F fittings used outdoors; however, alternate weatherproofing methods can be used.

I've found that adequate weatherproofing is assured by tightly wrapping the connectors and an inch of the cable with electrician's tape. Only a quality brand of PVC plastic electrical tape should be used; inferior imported or off-brands must be avoided.

The tape must be applied to a clean, dry surface. The tape will not stick properly in cold weather. In cold weather a silicone-filled boot is the best answer.

power-handling capability

The type-F fitting was designed for CATV/MATV without transmitting in mind. The type-F fitting will not degrade the rf-power rating, since spacings and dielectrics are the same as the cable. With certain chassis and adapter fittings, the spacings and dielectrics may differ since they are not made with cable. A derating of 50% of the *maximum* power rating for the cable should be adequate if you don't like to experiment.

For ac/dc current ratings, manufacturers of CATV/MATV equipment don't generally allow a type-F fitting to handle much over 1½ amperes. The experimenter will find that he will have no problem running higher currents, sometimes as high as 6 amperes.

salvaging used fittings

Type-F connectors can be salvaged without damage. Simply cut away the old ferrule with diagonal pliers and reassemble using a new ferrule. If a new ferrule is unavailable, a suitable one can be made from a piece of copper tubing.

assembly tools

With a little care an F fitting can be assembled with a pair of household pliers. For a good crimp, a crimping tool is suggested. This tool crimps the ferrule neatly in two places with less danger of splitting the ferrule. Heavy-duty and light-duty types are available. The parallel jaw tool is also very handy for miscellaneous jobs around the shack; for example, as a hand-vise or for straightening parts. A crimping tool is worth the extra cost if much crimping is done.

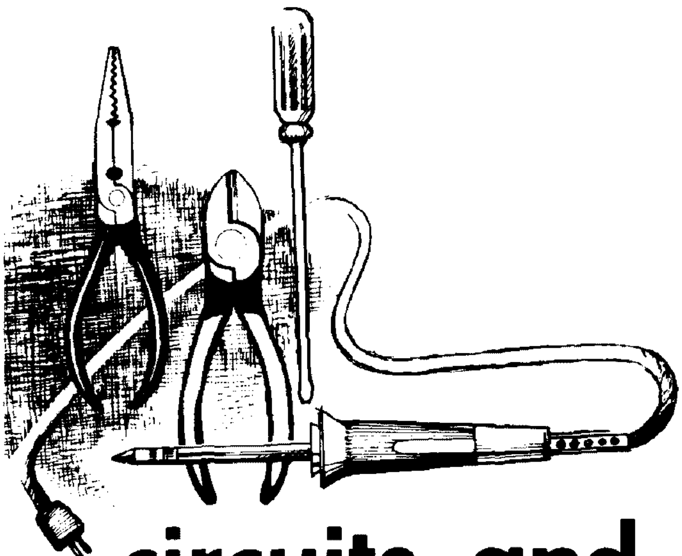
conclusion

It is my observation that many hams who are also involved in CATV/MATV work find many uses for type-F fittings. Most of my test equipment and ham gear has been converted to type-F fittings. Standardizing on this one fitting saves a lot of time and trouble hunting up various adapters and different cables.

acknowledgement

I would like to thank the following persons for their help in providing catalog information and technical assistance: George S. Bahue, Blonder-Tongue Inc.; Dick Mapes, Channel Master Corp.; Jim Emerson, Jerrold Electronics Corp.; Jerry Balash, JFD Electronics Corp.; W. E. St. Vrain, Mosley Electronics Inc.; and Hal Sorenson, Winegard Company.

ham radio



circuits and techniques

ed noll, W3FQJ

anti-QRM methods

Take note of your specific QRM problems today. Condemnations and groans are no help. Furthermore, QRM is likely to become more severe over the next several years. The short-skip, long-skip combination will prevail on 15 and 20 meters with long-skip signals diminishing and becoming more vulnerable to short-skip QRM. Twenty meters will fill to the brim as DX conditions fail on 10 and 15.

There will be a steady migration of DXers to the low-frequency bands. Ham population will increase, perhaps at a much faster rate outside of the United States. Night conditions on 40 and 80 will be as trying as ever; even 160 meters is bound to experience crowding. Is it a hopeless situation? It doesn't have to be!

In addition to those important essentials of gentlemanly operating practice and using no more power than necessary to maintain proper communications, each of us can take a number of steps to alleviate the severe QRM problem. Some are operational; others, technical.

operational considerations

There are a number of things we can do that will encourage all-band occupancy and the efficient and effective use of each band.

1. Use vhf for local rag-chewing and some of the more localized net activities.

2. Press for the legal restoration of the entire 160-meter band.

3. Do not give up so easily on the DX possibilities for 10 and 15 during minimum sunspot. Look for some new breakthroughs. Transequatorial (north-south) TE propagation[1] by way of chordal hops is a possibility for 6, 10 and 15 meters. Perhaps special emphasis should be given to 10 meters during the upcoming sunspot minimum. Massive cooperation among North, Central and South American hams could well uncover some unusual possibilities.

4. Continue with experimentation on unusual propagation paths on vhf/uhf, 10- and 15-meter bands. Are there any steps that can be taken through IARU that could implement satellite tests by radio hams worldwide?

5. In the USA we might set aside some spot frequencies to be retained by gentlemanly agreement for special technical activities. Such would be a boon to QRP experimentation, slow-scan television, low-frequency DX propagation tests and others.

The above activities and others as well would spread amateur activities more uniformly among the various bands and would make a major contribution toward QRM reduction on overcrowded frequencies.

There are technical steps that each of us can take to minimize or eliminate some of our own QRM problems.

1. Use a separate receiving antenna[2] or, perhaps a means of switching quickly among various antennas (including the transmit antenna) to find the one that will minimize a given QRM situation.

2. Use more directional antennas on 40, 80 and 160 meters. A directional receiving antenna is particularly helpful when QRM is bad.

3. The 10/15/20-meter directional beams as we know them today may not be the ultimate answer, especially as receiving antennas, in the medium and low sunspot years. The back pickup can be quite substantial and this is often in the short-skip direction when communicating with a long-skip station. The various phased arrays, although their forward gain may not be as high, can be designed with back direction nulls. A receiving antenna with an adjustable null point would aid in tuning out a specific piece of QRM.

4. There is a decade of amateur experimentation to be done on antennas with means of changing polarization and vertical angle to match propagation conditions of the moment.

5. Rotating loops function well on 20, 40, 80 and 160 meters for receiving. They may have some utility even on 15 and 10. A well-balanced shielded loop (circular or ferrite) has two deep nulls which can be used to tune out QRM. Sensitivity is often quite unimportant for routine operations on 40, 80 and 160.

6. Miniature receiving antennas such as the coil-loaded dipole or helix show great possibility for crowded strong-signal conditions.

7. The performance of loops and small antennas can be enhanced with the use of antenna-mounted preamplifiers, or preselectors positioned ahead of the receiver.

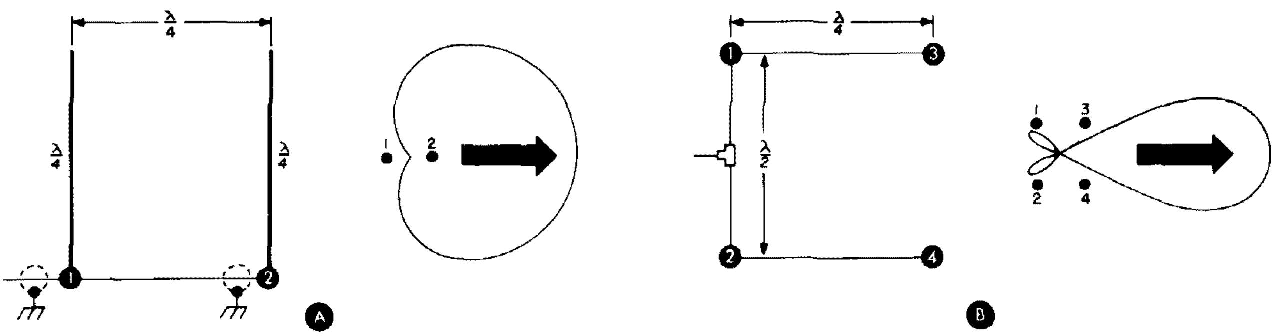


fig. 1. Layouts and patterns of phased quarter-wave verticals.

8. Tuners, input attenuators and couplers often improve receiving signal-to-noise and signal-to-interference ratios. If your receiver is prone to intermodulation distortion a well designed rf preamplifier and attenuator can cut back on splatter.

the transceive antenna

The antenna used for both transmit and receive modes is limited in its capabilities for establishing optimum receive conditions. Emphasis is on the most efficient transfer of power to the antenna and maximum radiation in a given direction. Some means for switching receiver input to one or more additional antennas even of a very simple type is a boon in reducing QRM. This applies in general to all bands, even in some cases for DX reception. It is a particularly useful capability for 40, 80 and 160 meters. In fact, on 80 and 160 the DX enthusiasts use separate transmit and receive an-

tennas to reduce QRM successfully. Switching between horizontally and vertically polarized antennas is common practice on 160. Quite often on 80 meters high horizontal transmit antennas are used along with a vertical receiving antenna. In many installations of this type either antenna can also be used for transmit.

Also popular are two phased verticals.[3,4,5,6,7] Usually a pattern switching facility is included. Verticals as effective transmit and receive antennas have not been fully exploited on the 10-, 15- and 20-meter bands. A properly fed and matched array of verticals can result in a very pronounced drop in pickup and radiation from the back. The back sensitivity of an antenna and lost power that is radiated to the rear are two causes of long-skip, short-skip QRM.

Two quarter-wave verticals spaced 90° and fed 90°-related, result in the attractive cardioid pattern in **fig. 1A.** Two such 90° pairs spaced 180°, **fig. 1B,** sharpen the forward pattern.[3,4] A suitable switching arrangement permits easy pattern changing whenever you want to enhance performance in another direction. Three verticals of this type spaced a quarter wavelength in a triangle permit complete rotation of the directional pattern (**fig. 2**).

Of course, QRM can also arise from the same direction as the desired signal with a short-skip station playing havoc with a long-skip signal. The big question is how much can be done to minimize this problem with a controlled vertical-radiation pattern.

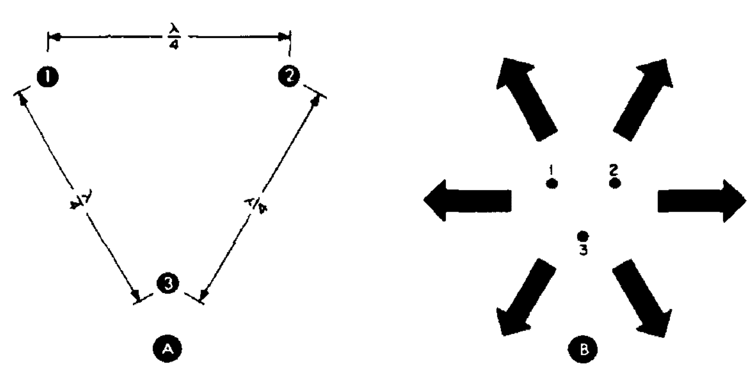


fig. 2. Three vertical antennas in a triangle (A), omnidirectional pattern (B).

separate receiving antenna

In an era of high sensitivity receivers the separate receiving antenna provides an effective weapon for combating QRM. Even a low horizontal dipole (**fig. 3**), rotated by a low-cost tv rotator, is surprisingly effective. You only have to aim the dipole elements in the direction of the interference to make use of its rather effective null. When receiving a reasonable signal, even a DX one, the antenna can be turned away from the desired bearing and still deliver a useful signal to the input of a sensitive receiver. If your receiver lacks sensitivity use a preselector ahead of its antenna input. It is common experience that an S3 to S4 signal in the clear is easier to read than an S7 to S9 signal clobbered by over-riding QRM. Full length dipoles can be used on 10, 15 or 20 meters; for three-band operation use a dipole with loading coils.

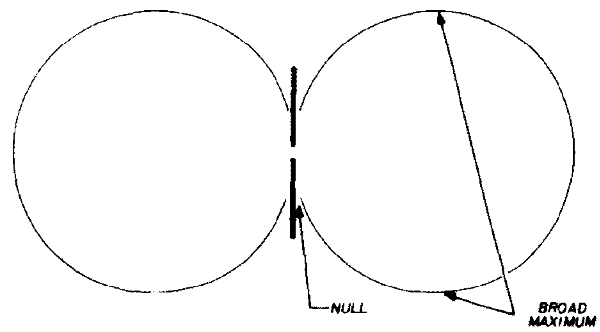


fig. 3. The dipole as a receiving antenna.

The quarter-wave vertical receiving antenna, loaded or full length, is popular on 40, 80 and 160 meters for DXing.

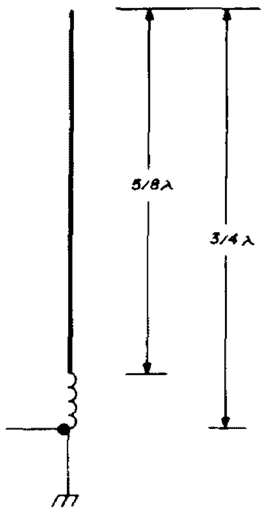


fig. 4. The 5/8-wave vertical antenna.

Such a vertical is less sensitive to high-angle signals with a consequent reduction in QRM. Two or three verticals provide directional horizontal patterns as well and a further reduction in QRM level.

Phased receiving verticals, ground mounted with suitable ground planes, do well as receiving antennas for 10, 15 and 20. Although a high beam may help you punch a stronger signal into a given

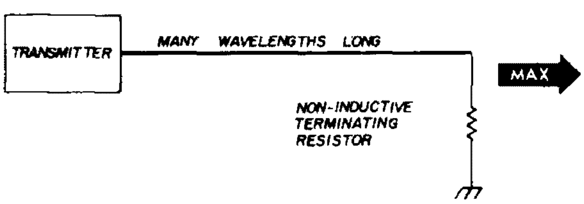


fig. 5. Beverage antenna.

corner of the world, the lower and limited-gain vertical array will often permit you to sort out the desired received signal from a maze of QRM.

The 5/8-wavelength vertical with its lower vertical angle has not been evaluated extensively in phased systems.[5] This may well be one of the finest receiving antenna systems available. A low impedance at the feed point is established with a loading coil that electrically tunes the vertical to 3/4 wavelength (**fig. 4**). A 5/8 wavelength vertical is approximately 20 feet long on 10, 30 feet on 15 and 40 feet on 20.

Long-wire antennas such as the Beverage[8,9] have demonstrated excellent possibilities for receiving from a single direction. W1BB and others have used them successfully in receiving 160-meter DX signals from Europe when other receiver arrangements were troubled with noise and QRM. The Beverage is a low long-wire antenna with its far end terminated in a non-inductive resistor. The antenna displays maximum receive sensitivity in this direction (**fig. 5**).

miniature receiving antennas

The short tuned antenna can be made to have rather deep nulls; at the same time the overall pattern is quite broad, either a figure-8 or cardioid. In heavy QRM the antenna can be rotated and the null positioned in the direction of the interference. In addition the mini antenna with its lower sensitivity cuts back on noise and intermodulation problems if they are a problem with your receiver.

A small loop antenna can be built for 40, 80 and 160 that gives you a good directional receiving pattern.[10,11,12] Loops can be used on 10, 15 and 20 meters as well. They are a particular help during strong short-skip conditions. In fact, the strong, short-skip conditions on 20 meters during the evening don't differ too greatly from those on 40. A receiving antenna of this type can be helpful for DXing when bands are cluttered with QRM. Although loops do not have the

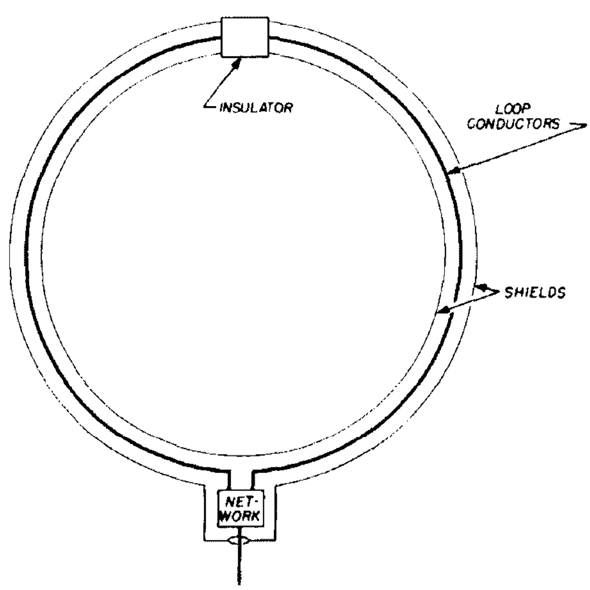


fig. 6. Basic loop antenna.

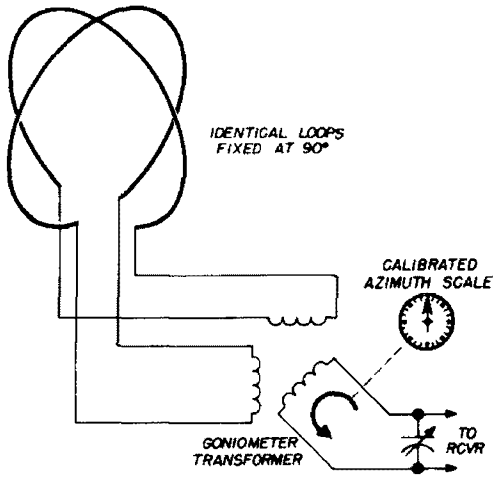


fig. 7. Bellini-Tosi fixed-loop goniometer.

sensitivity of your big antenna their deep nulls can often help you tune out QRM. The loop antenna and its transmission line should be kept well isolated from other antennas and lines to obtain a deep pattern null.

There are three basic types of miniature antennas: loop, ferrite core and helical dipole. Robert L. Nelson, K6ZGQ, has set down several rules of thumb for planning such an installation.[2] A practical level of radiation resistance and antenna efficiency for ham operation is obtained when the linear dipole-type antenna is longer than about 4 feet. The area enclosed by a loop configuration should be no greater than about 2 square feet.

Such antennas have been covered extensively in the literature as receiving loops and direction-finding antennas.[13,14] The much discussed Army loop is a larger specialized version of the same design. It is, of course, not as directional as a well built and shielded loop antenna.

Typical 160-meter loop diameter is two to three feet and consists of three or

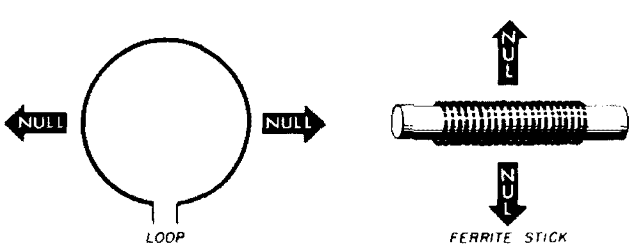


fig. 8. Nulls of loop antenna and ferrite loopstick.

four turns of insulated wire. A capacitor or inductor-capacitor network is used to resonate the loop as shown in **fig. 6.**

The loop coil is surrounded by a conducting shield which is connected to ground. The shield is broken at the top to provide an electrical opening for the signal. Soft drawn copper tubing can be used as the shield.

There are various other loop configurations worthy of consideration. Two loops

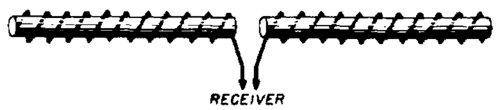


fig. 10. Helical dipole.

mounted near each other and properly phased can be used to obtain a unidirectional pattern with deep nulls. Two loops mounted at right angles to each other and using a Bellini-Tosi goniometer provide an electronically rotatable antenna pattern of the same type obtained with a single rotatable loop (**fig. 7**).

Many modern direction finders use a ferrite core and a tuned winding as in **fig.**

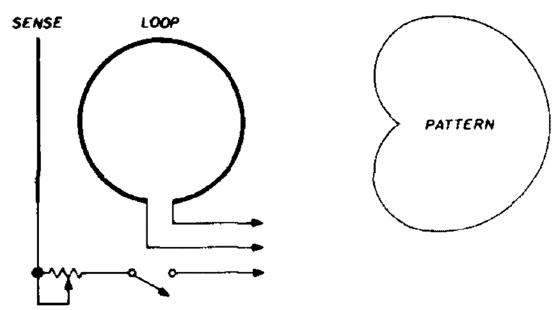


fig. 9. Loop with sense antenna produces cardioid pattern.

8. A typical ferrite directional antenna for the 2- to 8-MHz region consists of 14 turns of number-18 wire wound on a 3/8 by 6-inch ferrite loopstick.[14] It is tuned by a 200-pF variable. The nulls of a ferrite loop are perpendicular to the core (**fig. 8**). The axis of the core is in the direction of maximum sensitivity.

A short tuned vertical can be used

with a directional loop or ferrite stick to obtain a more unidirectional pattern as in fig. 9. (Its action is similar to the sense antenna of a direction-finding combination of loop and sense.) Some sort of attenuator is a part of the vertical, a necessity for equalizing the signal levels picked up by the loop and vertical. When signal currents are equalized, switching in the vertical creates a deep null toward the rear.

antenna can be used with a tuner, **fig. 11**, to extend its bandwidth if it is to be used as a transmit antenna.

An even greater length reduction can be obtained with end loading, as suggested by K6ZGQ[2] (fig. **12**). Capacitive hats, about 3 feet in diameter, can be made of stiff wire. Antennas no longer than 6 feet can provide operation on bands 20 through 160 meters; only the number of helical turns have to be changed. Two of these antennas can be mounted a short distance apart (approximately 0.05 wavelength) and fed out of phase. This produces a more unidirectional pattern and less pickup off the back.

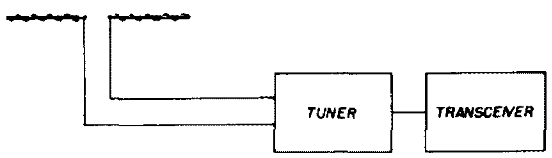


fig. 11. Helical dipole for transmit and receive requires tuner.

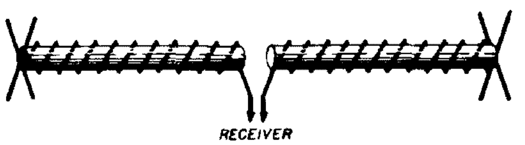


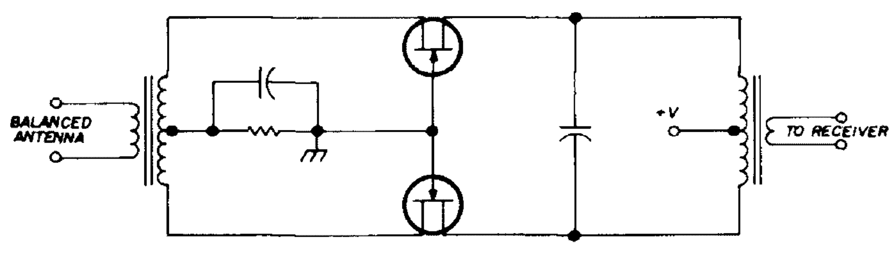
fig. 12. Ended-loaded helical dipole reduces overall length.

A helically-wound dipole is attractive for miniature receiving-antenna construction because of its horizontal polarization. Horizontal polarization is advantageous for those areas where there is a high level of vertically-polarized manmade noise. Two helical windings on each side of a dipole feed point, fig. **10**, permit a resonant antenna system with a short linear length. An advantage of the helical configuration is that antenna resistance does not differ much from that of a full-sized dipole unless the antenna is made extremely small. It can be used for both transmit and receive. Complete data for building helical antennas is given by W2EEY.[15]

receiver input systems

When using miniature and low-gain antenna systems that emphasize pattern rather than gain, additional amplification can be helpful. Antenna-mounted amplifiers have been used successfully in fm and tv reception and in other specialized receiving systems. Field-effect transistors and certain linear integrated circuits appear to be naturals for this application.

fig. 13. Common-gate dual fet amplifier is suitable for antenna-mounted amplifiers.



A linear helix with a length reduction factor of 0.3 (length 1/3 the length of a full-size dipole) usually provides practical operation over the cw or phone portions of a given low-frequency band. This

The dual jfet[16] and dual-gate mosfet[17] are particularly attractive because of low intermodulation distortion and their ability to handle a considerable range of signal levels without distortion.

When balanced feedlines are used a dual fet can be connected in a push-pull common-gate configuration, **fig. 13.** No neutralization is required.

Amplifiers of this type can be mounted in small weatherproof containers mounted at the antenna terminals. The dc supply voltage can be fed over the same coaxial cable that brings the signal down to the receiver. Such an installation can help substantially in building up the signal picked up by a highly directional but low-gain miniature antenna.

Antenna-mounted amplifiers for phased vertical systems are another possibility. The output of each individual vertical, **fig. 14,** can be supplied to a receiver-located tuner/phaser combination. The directional pattern of the antenna system can be oriented by controlling the relative phase and amplitude of the signals delivered by the vertical receiving antennas.

The integrated circuit is also attractive for antenna-mounted amplifiers. A case in point is the RCA CA3028A,[18] **fig. 15.**

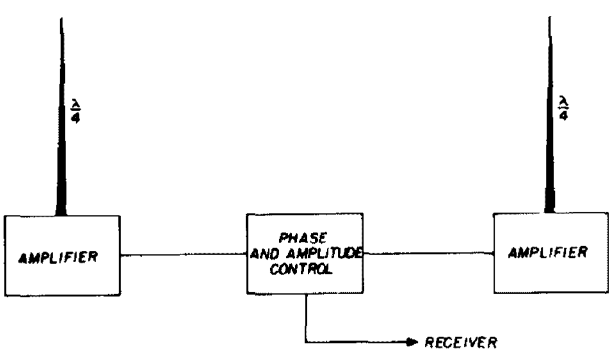


fig. 14. Phased verticals with amplifier and remote pattern control.

It consists of a single-stage differential amplifier, a constant-current transistor and suitable biasing resistors. Only a few external components are needed to build up an amplifier, **fig. 16.**

The same IC can also be operated as a converter. The implication here is that i-f conversion can be made at the antenna terminals. This early conversion to the i-f frequency can be helpful in reducing intermodulation distortion. The converter could possibly be tuned with a dc controlled voltage-variable capacitor diode.

Perhaps the ultimate in antenna-mounted amplification is the recently

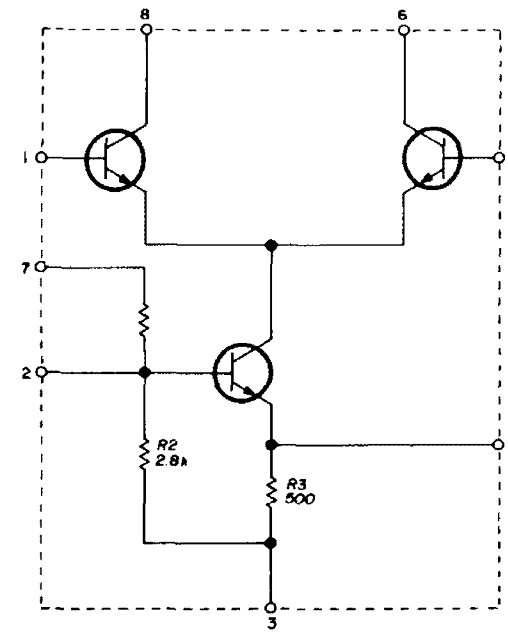


fig. 15. Internal circuit of the RCA CA3028 integrated circuit.

developed voltage-probe receiving antenna.[19] The antenna proper is a 2-inch rod loaded with a small disc and followed by a three-stage fet amplifier consisting of a low-noise input stage, driver and output stage. It will out-perform a 13-foot, 4-inch whip antenna from 7.5 MHz down and a 3-foot whip at all frequencies below 40 MHz.

There is a definite trend in the design of modern amateur communications receivers toward the thorough reduction of intermodulation distortion.[20,21,22,23,24,25] Even the best commercial receivers are often not able to cope with the wide range of incoming signal levels, intermod and splatter. A good agc system and fet input are helpful. There are great possibilities for the direct-conversion receiver and other designs that have low noise content, wide dynamic range and low distortion. An assist to these objectives is an input system of relatively low gain ahead of the second

mixer. A low-noise first mixer and a quiet low-noise i-f chain are important. An attenuator ahead of the receiver input, with or without tuner, can help to reduce intermodulation distortion.

If you suspect noise level and intermodulation distortion in your receiver is affecting sensitivity try a good low-noise preselector (with wide dynamic range) ahead of your receiver.

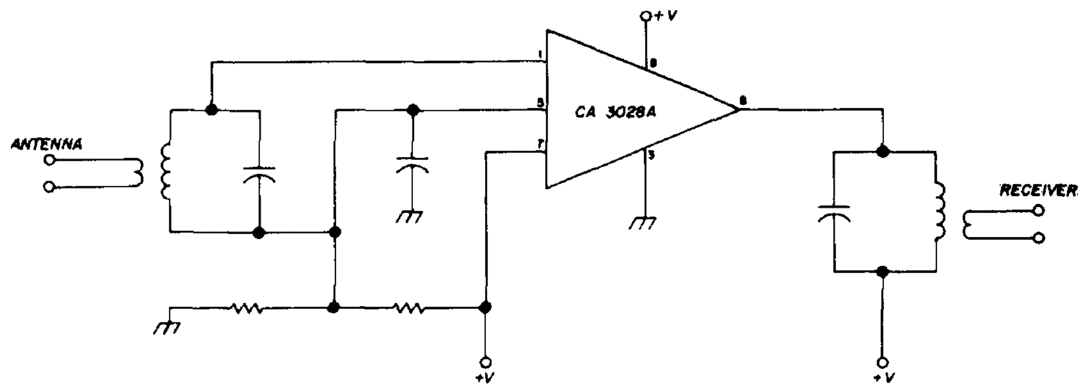
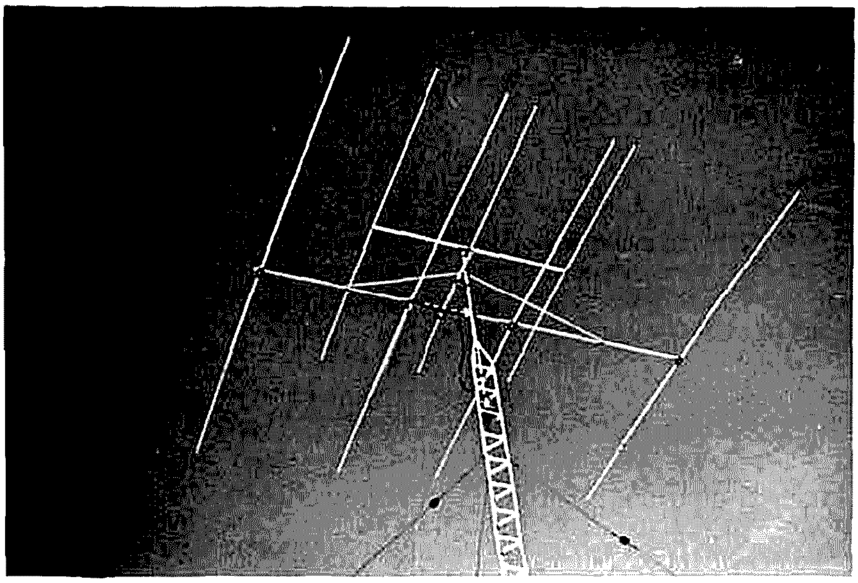


fig. 16. Using the CA3028 ic as an i-f amplifier.

bibliography

1. Pat Hawker, G3VA, "Transequatorial Propagation," *Radio Communication*, October, 1970, page 682.
2. Robert L. Nelson, K6ZGQ, "Receiving Antennas," *ham radio*, May, 1970, page 57.
3. George Cousins, VE1TG, "80-Meter Vertical Beam Antenna," *ham radio*, May, 1970, page 26.
4. Malcolm M. Bibby, GW3NJY, "Uni-Directional Antenna for the Low-Frequency Bands," *ham radio*, January, 1970, page 61.
5. Edward M. Noll, "73 Vertical, Beam and Triangle Antennas," Howard Sams Co., page 7.
6. Edward M. Noll, Phased Verticals, "73 Vertical, Beam and Triangle Antennas," Howard Sams Co., page 37.
7. Pat Hawker, G3VA, "Directional Verticals," *Radio Communication*, September, 1970, page 605.
8. Pat Hawker, G3VA, "Beverage Aerials," *Radio Communication*, October, 1970, page 685.
9. John D. Kraus, Beverage or Wave Antenna, "Antennas," First Edition, McGraw Hill Book Co., Page 412.
10. Richard A. Genaille, W4VW, "A Top Band Loop Antenna," *CQ*, January, 1969, page 41.
11. A. L. Bailey, G3WPO, "Loop Aerial for Top Band," *Short Wave Magazine*, October, 1969, page 482.
12. Spenny, G6NA, "Loop Aerials," *Radio Communication*, September, 1968, page 576.
13. Robin J. Pearce-Boby, G3JLE, "Direction Finding and D/F Receivers," *Radio Communication*, April, 1969, page 238.
14. Sam Kelly, W6JTT, "Solid-State Radio Direction Finder," *ham radio*, March, 1970, page 38.
15. John J. Schultz, W2EEY, "Constructing Efficient Helical Antennas," *CQ*, May, 1968, page 34.
16. Ralph W. Campbell, W4KAE, "Differential J-FET Preamplifiers," *73*, November, 1970, page 16.
17. Suleyman Sir, "RF Applications of the Dual-gate MOSFET," *73*, November, 1970, page 22.
18. H. M. Kleinman, "Application of the RCA-CA3028A and RCA-CA3028B Integrated-Circuit RF Amplifiers in the HF and VHF Ranges," RCA Application Note ICAN-5337.
19. James R. Fisk, W1DTY, "Voltage-Probe Antenna," *ham radio*, October, 1970, page 20.
20. Pat Hawker, G3VA, "Solid-State Receiver Design," *Radio Communication*, October, 1970, page 683.
21. F. H. Littlefield, Jr., K1BQT, "Direct-Conversion Receiver," *ham radio*, October, 1970, page 16.
22. William Sabin, WOIYH, "The Solid-State Receiver," *QST*, July, 1970, page 35.
23. Pat Hawker, G3VA, "More on Narrow-Band Reception," *Radio Communication*, October, 1968, page 651.
24. Ham Notebook, *ham radio*, January, 1970, page 70.
25. "Receiving Aerials," RSGB Radio Communication Handbook, Fourth Edition, page 1,380.

ham radio



single feedline

for multiple antennas

A decoupling scheme for exciting separate antennas with one transmission line

D. W. Bramer, K2ISP, 125 Winfield Road, Rochester, New York 14622

Many articles have appeared that discuss the merits of single-band rotary beams versus those featuring multiband operation. One popular solution to multiband antenna operation is the tribander. This approach has two disadvantages: Element spacing can be optimized for only one of the three bands, and resonant traps and impedance-balancing devices introduce ohmic losses. Both effects can't help but compromise overall antenna efficiency.

Another solution to multiband antenna operation is the log periodic array. While theoretically the most satisfying solution, the log periodic antenna isn't too practical for most amateurs because of cost and structural complexity. In addition, the size of the log periodic antenna diminishes in its effect as frequency increases, while its gain remains essentially constant.

Still another multiband antenna approach uses interlaced elements on a single boom. This scheme permits more optimal element lengths, reduced ohmic losses, and proper spacing with minimal element interaction. Still, the interlaced beam is tricky to adjust and clumsy to install.

The only real solution to effective multiband antenna operation is to use a separate antenna optimized for each favorite band. This implies the use of separate feedlines for each antenna, or complex remote-control circuits for transferring a single transmission line between antennas.

This article describes an efficient method of feeding separate antennas on two or more bands with a single coax cable using a well-known principle of transmission-line theory. Superior antenna performance can be obtained at moderate cost.

economic considerations

A point often overlooked is that you can purchase three separate antennas of excellent mechanical and electrical quality for nearly the same price as a 20-,

15-, and 10-meter tribander. In fact, when considering only 10 and 15 meters, you can own separate 4-element yagis on each band for just a little more than the list price of one of the new 3-element dual-band antennas. Thus for a comparable investment, it's quite easy to achieve far superior performance in terms of measurable forward gain, directivity, and f/b ratio together with an exact impedance match; not to mention the absence of weather-susceptible traps. Moreover, such antennas can be home constructed using standard catalog boom-and-mast hardware for even greater savings.

At the risk of being accused of beating the cost issue to death, I'd like to make one more point. Consider that the price of a couple hundred feet of *good quality* coax with connectors comes quite close to that of a good commercial 3-element 10-meter beam. If you add the relays, control wiring, and switches necessary for separate antenna selection with a single feedline, the feed-system cost will be even higher.

the gimmick

A review of basic theory shows that, in the case of a half-wavelength transmission line, voltage and current are identical at the input and output terminals of the line. Hence the input impedance of any line, regardless of its characteristic impedance, is exactly the same as the load impedance, providing the line length is an *exact multiple of a half wavelength.* Such a line, therefore, may be used to transfer an impedance to a new physical location without changing the line's intrinsic characteristics. A typical application of this principle to achieve common-line feed for two different single-band antennas is shown in fig. 1.

Assume that each antenna has been individually adjusted to reflect 52 ohms pure resistive load at each respective nominal operating frequency, f_c. The 10-meter yagi is connected in parallel with the 15-meter yagi; hence it is also connected to the main feedline via a short length of coax. This interconnecting piece acts as our simple decoupling gimmick, which is electrically one-half wavelength long at $f_c = 21.3$ MHz of the lower-frequency antenna.

At 21 MHz the 10-meter yagi reflects a complex load consisting of $X_C + R$, with a net impedance much higher than 52 ohms, because its resonant frequency is

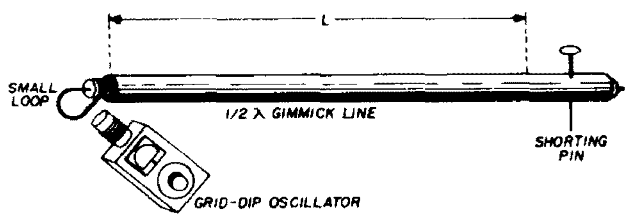


fig. 2. Test setup for determining appropriate length of the decoupling gimmick. Length, L, is trimmed for the operating frequency of the lower-frequency antenna.

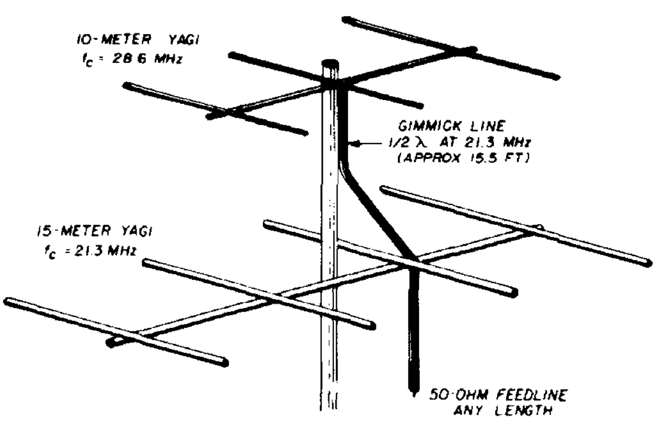


fig. 1. Decoupling method for isolating two stacked beam antennas. The 1/2-wave gimmick may be coiled into a loop and fastened to the mast or to the boom of the higher-frequency array.

far removed from 21 MHz. This same net high impedance is repeated one-half wavelength away (at 21.3 MHz) across the input terminals of the 15-meter yagi. The 15-meter beam reflects a much lower load (52 ohms) to the main transmission line; so the 15-meter beam absorbs power, while the 10-meter beam accepts essentially no power.

At 28 MHz, the 15-meter yagi reflects the much higher net impedance consisting of $X_L + R$. When operating on 10 meters the 10-meter yagi absorbs power, while the 15-meter yagi accepts virtually no power. The interconnecting gimmick line

looks simply like a short continuation of the main transmission line.

construction and installation

Construction of a decoupling gimmick is very simple, as it is nothing more than a piece of coax trimmed to a fairly critical length. As a start, consider overall length to include all connections; i. e., coax fittings, terminal lugs, etc. Cut a piece slightly to the high side in length using:

$$L = \frac{500}{f_{MHz}} \times \text{velocity factor}$$

where $f_{MHz} = f_c$ of lower-frequency antenna

Install appropriate terminals or coax fittings to one end only. At this same end, temporarily provide a small loop, either by bolting terminals together or by fastening a ¾-inch-diameter loop to a coax fitting. The setup is shown in **fig. 2.** Short-circuit the opposite end by driving a large pin or nail through both shield and inner conductor. Using a grid-dip oscillator with a calibrated receiver, determine the resonant frequency of the initial length of cable. Then change the position of the shorting pin in gradual steps until the desired f_c is obtained. Cut off the excess cable, install appropriate fittings, and your decoupling gimmick is complete.

If your antennas are spaced closer together than the overall decoupling gimmick length, form the excess cable into a coil about 8 inches in diameter. Tape the coil to the mast or to the boom of the higher-frequency antenna.

experimental data

My installation, consisting of 10- and 15-meter gamma-matched beams, is shown in the photograph. The antennas are stacked 5 feet apart and consist of 3 elements on a 10-foot boom for ten meters and 4 elements on a 21-foot boom for 15 meters. Several experiments resulted in the following data.

Each antenna was tuned individually while completely divorced physically from the other. The data, shown in table

table 1. Measured data using home-made test equipment. The standing-wave ratio was unity at fc.

antenna	forward gain (dB)	fc (MHz)	f/b ratio (dB)	bandwidth for 1.75 swr (kHz)	horizontal beamwidth at 3-dB points (deg)
10-meter beam	8.1	28.6	23	950	50
15-meter beam	9.3	21.27	27	600	43

1, was acquired using a home-made swr bridge and field-strength meter.

Next, both antennas were mounted in final position on the mast. Alternately, one yagi was excited directly with the other completely disconnected. Rough data was taken as above, and results were essentially identical except that the 10-meter yagi was detuned; its resonant frequency was about 50 kHz higher.

The decoupling gimmick was installed, and the data in **table 1** was rechecked. Results were again identical, except that the 1.75:1 swr bandwidth decreased about 75 kHz for each antenna. This was considered to be well within usable limits of operation.

A reflected-power meter was inserted between antenna terminals and connecting lines. With moderate power applied, forward power to each *off-frequency* antenna was small compared with total transmitter output; in each case, reflected power was nearly equal to forward power. This confirmed that the decoupling scheme was indeed providing proper antenna isolation.

conclusion

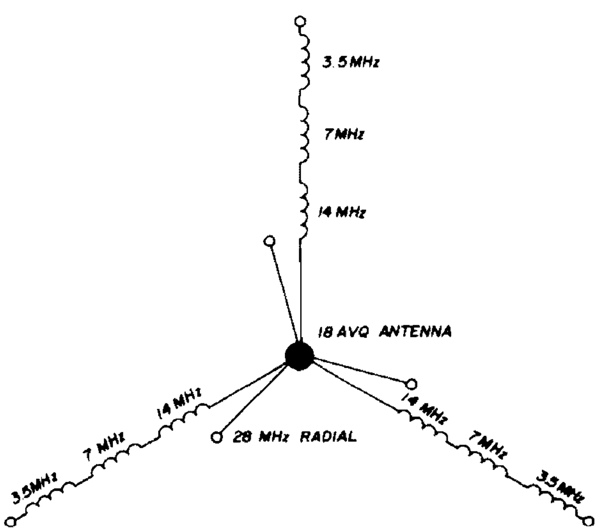
The system described has been in operation for nearly two years. Many notable DX stations around the world will attest to consistent signal punch, notable scores in DX competitions, and distinct recognition in rare-country pile-ups. This system has made significant contributions toward my list of confirmed countries in the ARRL DX Century Club listing.

ham radio

the ham notebook

multiband ground-plane

Although the popular Hy-gain 18AVQ multiband vertical antenna is designed to be mounted on the ground, it can be used as an elevated ground plane. All that is required is a suitable arrangement of inductance-loaded radials that provide resonance on each of the amateur bands.



radial loading coils

14 MHz	29 turns no. 20, 1.5-inch diameter, 16 turns per inch (Air-Dux 1216T)
7MHz	46 turns no. 18, 2-inch diameter, 16 turns per inch (Air-Dux 1616T)
3.5 MHz	94 turns no. 18, 2-inch diameter, 16 turns per inch (Air-Dux 1616T)

fig. 1. The Hy-Gain 18AVQ multiband vertical can be used as a ground-plane antenna by using a system of inductance-loaded radials.

JA1QIY reports excellent results with the system shown in **fig. 1**. The radials shown in **fig. 2** provide high performance on 3.5, 7, 14 and 21 MHz; a separate set

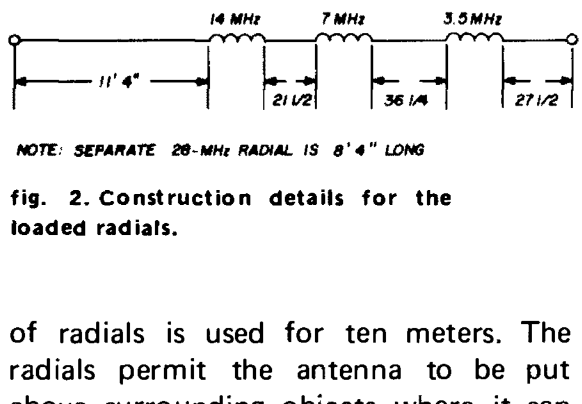


fig. 2. Construction details for the loaded radials.

of radials is used for ten meters. The radials permit the antenna to be put above surrounding objects where it can do the most good. If the radials are allowed to slope away from the 18AVQ they can also be used as guys. With the radials sloping away from the antenna at about 45°, the antenna provides a relatively good match to 50-ohm coaxial cable. The dimensions in **fig. 2** are for the cw end of the band, but with a little cut and try, equal performance can be obtained on the phone bands. The multiband swr of the JA1QIY antenna is shown in **fig. 3**.

JA1QIY has reported excellent DX performance with this antenna, particularly on 80 and 40 meters. On 80 he has worked Soviet Russia, Korea, Okinawa and the Philippine Islands; on 40 he has been able to work into the United States

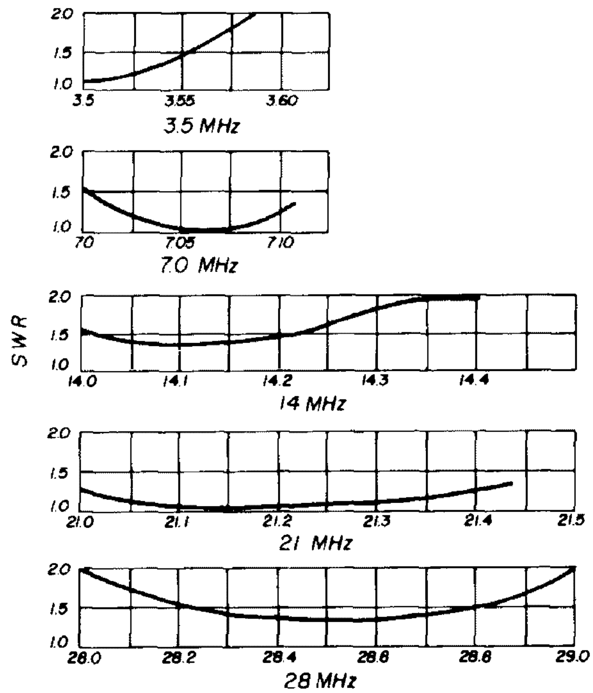


fig. 3. Swr performance of the multiband ground plane. Loaded radials were designed for the cw end of each band.

and Canada, no small feat with relatively low power.

In the original version of this antenna each of the radial loading coils is mounted around a length of phenolic rod. However, ceramic strain insulators could be used for more strength and better performance in wet weather. The 28-MHz radials are spaced a few inches away from the low-frequency radials.

multitester

Although most ham shacks have at least one volt-ohmmeter or vtvm, a low-cost utility multitester can fill in when you have to make simultaneous voltage and current measurements. The multitester in **fig. 4** covers the most useful voltage and current ranges and uses a low-cost 1-mA meter movement. The multiplier and shunt resistors, and diode are mounted on the selector switch, a Centralab PA1001. All resistors can be 5% carbon composition types, although for higher accuracy 1% precision resistors should be considered. However, to keep the cost down, carbon composition resistors provide acceptable accuracy for most purposes. R1, the 100-mA shunt, consists of 5-feet no. 36 wire wound around a high-value resistor. The 1-mA meter is an imported unit, such as the Lafayette 99F50528.*

*Available from Lafayette Radio Electronics, 111 Jericho Turnpike, Syosset, L. I., New York 11791. $2.95 plus postage; shipping weight, 12 ounces.

silver/silicone grease

A common problem with station ground systems is corrosion at the main ground connection. Some amateurs try non-conductive grease to eliminate this corrosion but this leads to improper grounding conditions. Silver/silicone grease is prepared by Technical Wire Products, Inc. (427 Olive Street, Santa Barbara, California 93101) for use on knife switches in power substations. However, it is useful to amateurs who want to protect their ground-system connections from corrosion and resultant loss of effectiveness. The silver/silicone grease is water repellent and is available in 2-ounce tubes (part number 72-00016) or 1-pound jars (part number 72-00015). For more details on this grease, write to Technical Wire Products, Inc. for a copy of Data Sheet CS-725.

Bill Welch, W6DDB

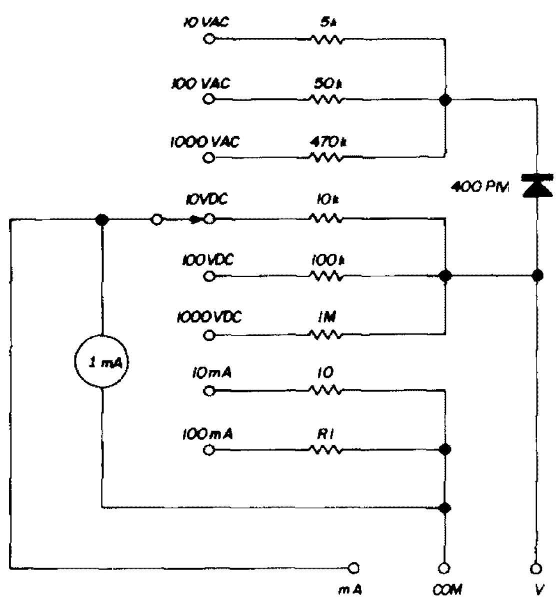


fig. 4. Simple utility meter. R1 is 5 feet no. 36 wire wound around high-value resistor.

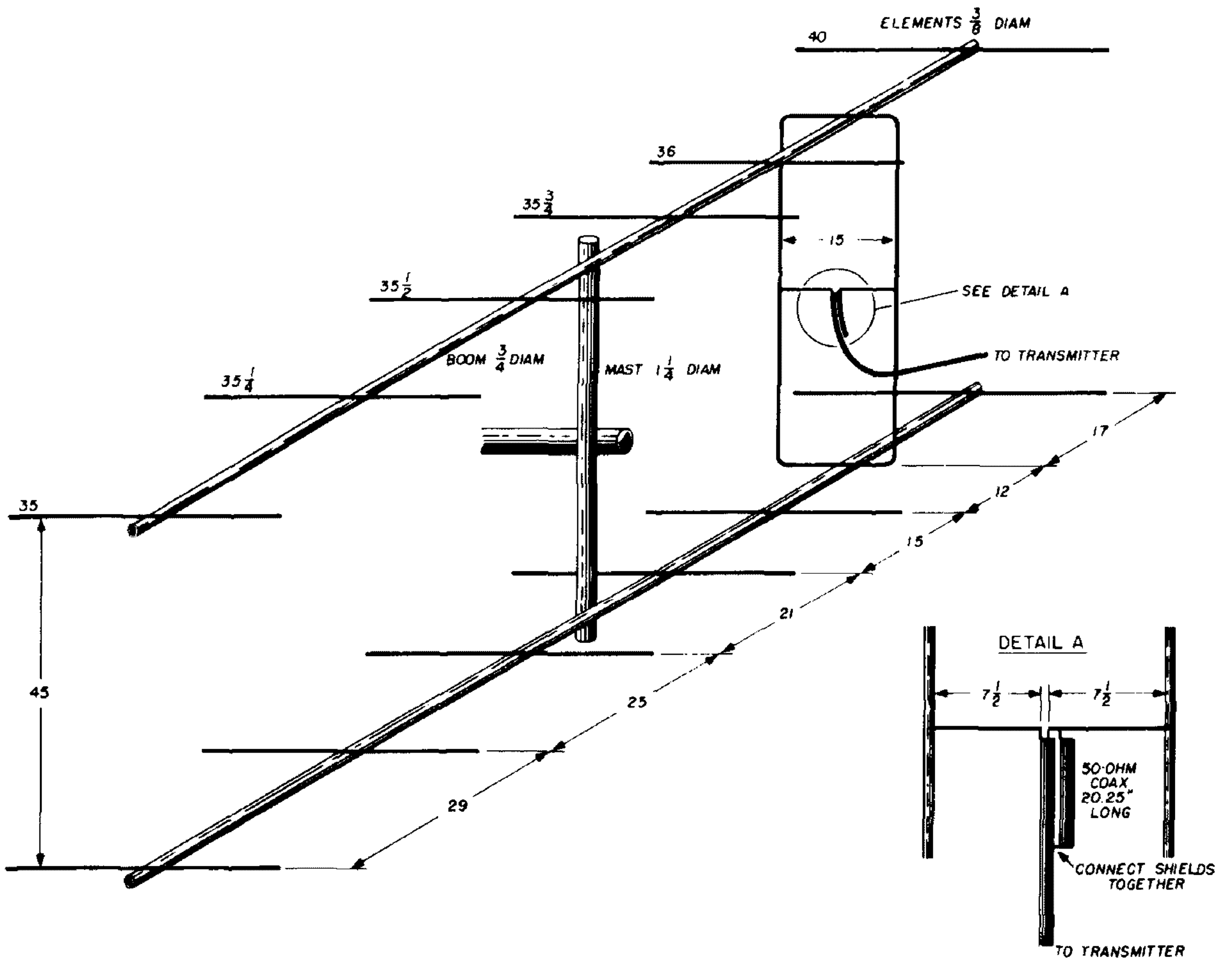


fig. 6. Details of two-meter slot antenna built from salvaged tv-antenna tubing and thin-wall conduit.

power-supply hum

The power supply described by Hank Olson in the February 1970 issue of *ham radio* was built for a solid-state receiver. The receiver was very quiet until the antenna was connected, then I found a wild example of "tunable hum." A 0.1 μF disc ceramic capacitor was connected as shown in fig. 5, and the trouble was cured completely. In my case it was convenient to connect the capacitor across one-half of the secondary, but it is equally effective across the entire secondary winding.

Bill Wildenhein, W8YFB

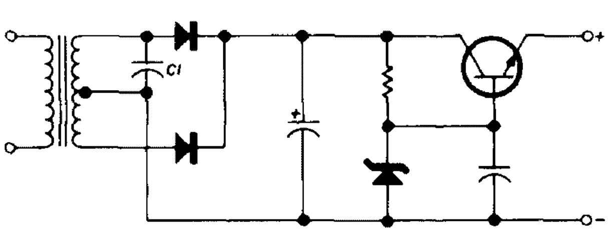


fig. 5. Capacitor C1 reduces power-supply hum.

two-meter fm antenna

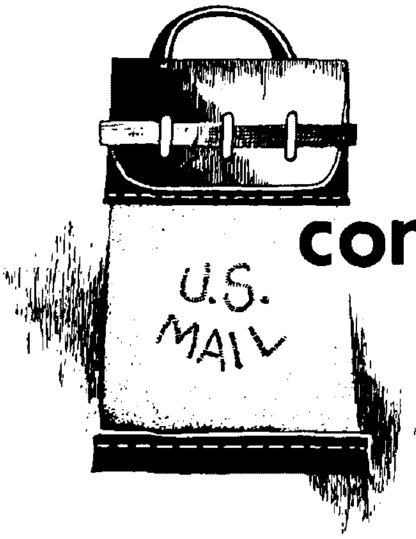
The antenna shown in fig. 6 was assembled from a few old TV antennas for about $3.00. My location is surrounded by mountains as high as 11,500 feet. The antenna shown has given good results by bouncing signals off the mountains. Stations 120 miles away sound almost like locals.

construction

The booms and crosspiece are made of 3/4-inch thin-wall conduit. All elements are 3/8-inch aluminum tubing salvaged from old TV antennas. The corners of the driven element are made of soft aluminum, assembled with ¼-inch bolts with the heads cut off.

I drilled holes in the conduit to accept the elements, which are held in place with sheet-metal screws. The booms are secured to the crosspiece with U-bolts at the balance point.

Bill Done, WB6KYE



comments

inverted-vee antennas

Dear HR:

In my experience with an inverted-vee antenna for 40 meters, 67.5 feet long with the apex at 30 feet and the ends 15 feet off the ground, resonant frequency was 7.18 MHz. This suggests that inverted-vee length is:

$$\text{length (feet)} = \frac{485}{f_{MHz}}$$

The resonant frequency was indicated by a very deep and sharp null with an Omega-T antenna noise bridge.

It may be of interest to comment on the affect of adding a parasitic inverted-vee director, 64.13-feet long (5% shorter than radiator), at the same apex and end heights, spaced 16 feet in front of the driven element. (Judging from Yagi data this is good spacing for gain, but not for bandwidth.) The director raised the resonant frequency 114 kHz to 7.294 MHz. This suggests:

$$\text{driven element (feet)} = \frac{492}{f_{MHz}}$$

$$\text{director length (feet)} = \frac{467}{f_{MHz}}$$

Element spacing was 0.11 wavelength.

The Omega-T noise bridge indicated a driving-point impedance of 27 ohms for this two-element inverted-vee Yagi. Two such antennas were used on 40-meter phone and cw during the 1950 ARRL Field Day with good success. A 38-ohm Q-section will match 27 ohms to 50-ohm coax, two 22.93-foot lengths of RG-59/U (length = 165/fMHz) were soldered in parallel to make a 37.5-ohm Q-section. According to the book, the Q-section should be made from twin lead, but with the double-coax Q-section swr was less than 1.4:1.

Another two-element inverted-vee Yagi with a director 3% shorter than the radiator resulted in the following formula:

$$\text{driven element (feet)} = \frac{490}{f_{MHz}}$$

$$\text{director length (feet)} = \frac{467}{f_{MHz}}$$

Element spacing for this array was also 0.11 wavelength.

Bob Hume, WB6AQF
Palos Verdes, California

ground rods

Dear HR:

Driving of 6-foot ground rods is easy in good loam, but many of us live in areas where the top soil is sandy or rocky, and you cannot find even a very poor ground with a 6- or 8-foot rod. It is necessary to get through hardpan or layers of rock to find moisture, to say nothing of water. I had to go down 18 feet.

Some small ground rods are made so they can be threaded together, but the

driving cap furnished with them comes loose with each blow of the sledge and buggers the threads. Also, when threaded they unscrew underground.

The same is true of well hardware. Ordinary ground points tend to pack the soil around them, and thus insulated themselves from actual ground. Well points will do the same, but due to their greater diameter, they pack less. A 1½-inch well point can be driven into soft soil, but the threads get mashed even when using the fitting made for driving.

The easiest way in the long run is to get a 1½-inch well point and several sections of extension pipe. Drill a ¼-inch hole at the top of one pipe and hang two buckets of rocks from the top. I used 5-gallon paint cans. Fill the buckets with rocks and then fill the leftover space with water for weight.

Cut off the end of an old 3/8-inch hose and drop the end down into the well point. Use a pair of chain tongs and take about three or four turns in the assembly with the water running easy. Then take half a turn back; keep the water running.

The pipe should run down about an inch for each 4 turns. After running it into the ground as far as necessary to reach the ground-water level fill the pipe with copper sulfate and stick a funnel into the top. It will catch enough rain to melt about 5 pounds of copper sulfate a year. A 25-pound sack of copper sulfate costs about $6.00 wholesale.

I finally ran my 18-foot ground in after only 10,000 turns or so!

Keith Olson, W7FS
Belfair, Washington

checking coaxial cable

Dear HR:

I wonder if any other amateurs have stumbled onto a very useful secondary function of my QRP wattmeter (April 1970 *"ham radio"*), *coaxial cable quality checking.* You simply feed the output of your QRP rig or exciter into the rf wattmeter. This will give a reading (let's say) of 7 on the meter. Next, the cable to be tested is inserted between the exciter and the wattmeter. If the cable has a nominal 50 ohms impedance the wattmeter reading should be reasonably close to the earlier reading. If not, the watts lost in a given length of 50-ohm cable can be directly calculated by subtracting Watts $_{far}$ from Watts $_{near}$ as read directly from the QRP wattmeter.

Neil Johnson, W2OLU
Tappan, New York 10983

fm deviation measurement

Dear HR:

Accurate fm deviation measurements are much more involved than they may seem. The i-f and discriminator must be extremely broad. Pulling out the i-f filter and reducing the size of the discriminator load resistors by about 3:1 is a good start. It's also a good idea to tweak the discriminator for best linearity — it's probably not as good as you might think.

In reference to the fm deviation meter described in the *ham notebook* section of the December, 1970 issue, a somewhat less expensive calibration procedure involves switching in a 455-kHz crystal first, than a 440-kHz or 470-kHz crystal in place of the second conversion oscillator crystal. If you don't have a dc-coupled oscilloscope, you can use the same technique used at the Motorola factory: use an audio-frequency multivibrator to switch between a 440-kHz and 470-kHz oscillator, and couple the output into the receiver i-f strip.

If the whole thing sounds a bit cumbersome, you can set your deviation quite adequately for amateur purposes by bellowing into the microphone, and adjusting the deviation control to just below the point where first limiter current starts to kick downward with modulation. This may not sound very sophisticated, but it's probably at least as accurate as most homemade deviation meters.

J. A. Murphy, K5ZBA
Tulsa, Oklahoma

new products

hallicrafters five-band transceiver

A new and improved version of the popular Hallicrafters SR-400, Cyclone II, ssb cw transceiver is being introduced. Known as the model SR-400A, Cyclone III, the new 5-band ssb cw transceiver is "feature-packed" with the newest and advanced electronic circuits and control designs developed for operation in all environments — field, radio shack or mobile. Full coverage is provided for the 80, 40, 20, 15 and 10 meter bands.

The ssb power of the new SR-400A has been increased to 550 watts PEP. An exclusive power-amplifier-tube balancing circuit eliminates the need for matched tubes. No longer does the amateur operator have to purchase final amplifier tubes in matched pairs.

Also, Hallicrafters has engineered the SR-400A with a new built-in 100/25 kHz crystal calibrator which helps locate band edges on the new sub-bands. The Hallicrafters' patented Receiver Incremental Tuning (RIT) permits ±3 kHz adjustment of receiver frequency independent of the transmitter is now calibrated, making it even more versatile and easier to use. Also available is an optional fixed station blower kit; it consists of a quiet, heavy-duty air blower and connections for use in the power amplifier section to prolong tube life.

The new SR-400A, like its predecessor, is ideal for the sophiscated DXer when used with the HA-20 remote vfo. The operator can listen to two frequencies simultaneously. It is also excellent for "tail ending." The HA-20's dial is calibrated for true 1 kHz readout. An expanded vswr meter with remote bridge connection allows the operator to continuously check antennas performance.

For the cw operator the SR-400A provides such added features as semi-automatic break in, cw filter (200 Hz), cw sidetone and grid-block keying. The SR-400A also features upper- and lower-sideband selection, constant tuning rate on all bands, smooth gear-driven tuning mechanisms with 1-kHz readout, choice of vox or ptt (built-in), 6-pole 1650-kHz crystal-lattice filter for optimum selectivity and sideband response, built-in notch filter to supplement sharp cw selectivity, frequency zero adjust, complete metering and separate receiver and transmitter rf and audio controls.

For the amateur who wishes to further extend the capability of the new SR-400A, Hallicrafters has provided optional accessories such as solid-state ac and dc power supplies, mobile mounting rack, and a 4-inch communications speaker for mobile installations.

The new SR-400A cyclone III weighs 26 pounds and is priced at $995 amateur net. The HA-20 vfo is priced at $199.95. The new SR-400A will be available from franchised electronic distributors throughout the United States. For more information, use *check-off* on page 94.

More Details? CHECK—OFF Page 94

callbook supplement

The *Radio Amateur Callbook* has announced an innovation in their line of publications. Beginning with the Spring 1971 *Callbook,* there will also be published a supplement, available to those subscribers who have purchased the preceding issue and would like to up-date their information without purchasing a complete book. Those amateurs who purchase the current edition may also order the supplement to the next complete edition for a modest cost. This is on an experimental basis, pending response from the amateur fraternity.

Supplements will contain all new licenses and silent keys, as well as call letter, class and address changes that have taken place since the preceding issue of the *Callbook.* For further information write The *Callbook,* Lake Bluff, Illinois 60044, or use *check-off* on page 94.

electronic designer's handbook

The "Electronic Designer's Handbook: A practical Guide to Transistor Circuit Design" by T. K. Hemingway is now in its second edition. This up-to-date handbook provides a complete reference on transistor circuit design to the depth required for practical engineering design. Part 1 provides detailed coverage of the transistor used as a switch and as a small-signal amplifier, as well as circuit operating principles and consideration of transistor parameters in practical design. The designer will find part 2 of particular value for its description of unusual circuits, and the straightforward discussion on how novel design can be synthesized and modified to serve in a number of practical applications. The content is specifically intended to show the reader how to design his own circuits. As opposed to presenting superficial knowledge of a great number of circuits, specific circuits are presented and analyzed in detail so that the reader, armed with the underlying design techniques, can apply

them to developing his own specific circuits.

Part 3 discusses the practical difficulties encountered in the design and testing of prototype circuits, offering many useful hints for solving a number of problems rarely mentioned in other books, and containing a particularly interesting and useful chapter on circuit layout and wiring. Much of the math has been relegated to the appendix. This book is specifically oriented to serve the needs of the practical designer, whether he be a novice or professional. 294 pages. 304 illustrations, $9.95 from Tab Books, Blue Ridge Summit, Pennsylvania, or use *check-off* on page 94.

automatic tone encoders

The new solid-state Alpha miniature continuous-silent-tone encoder is designed for use in all two-way radio communications systems and equipment. This compact, easily installed unit provides an end to interfering signals and the necessity of listening to co-channel users by requiring that transmissions to base station radios be accompanied by a predetermined subaudible tone to activate the receiver. The silent-tone encoder (ST-85H) can be used where the base station only is to be quieted. The new transistorized circuitry (eliminating mechanical reeds and relays) makes this thoroughly field-tested unit stable and reliable.

Tone frequencies are available over a wide range (20.0 to 203.5 Hz) and can be changed by simply plugging in a TN-91 of the frequency desired. Each silent-tone contains two sections, the encoder network (ST-85H), and the tone determining network (TN-91H), and features small size, built-in voltage regulation and one year warranty.

For more information write to Alpha Electronic Services Inc., 8431 Monroe Avenue, Stanton, California 90680, or use *check-off* on page 94.

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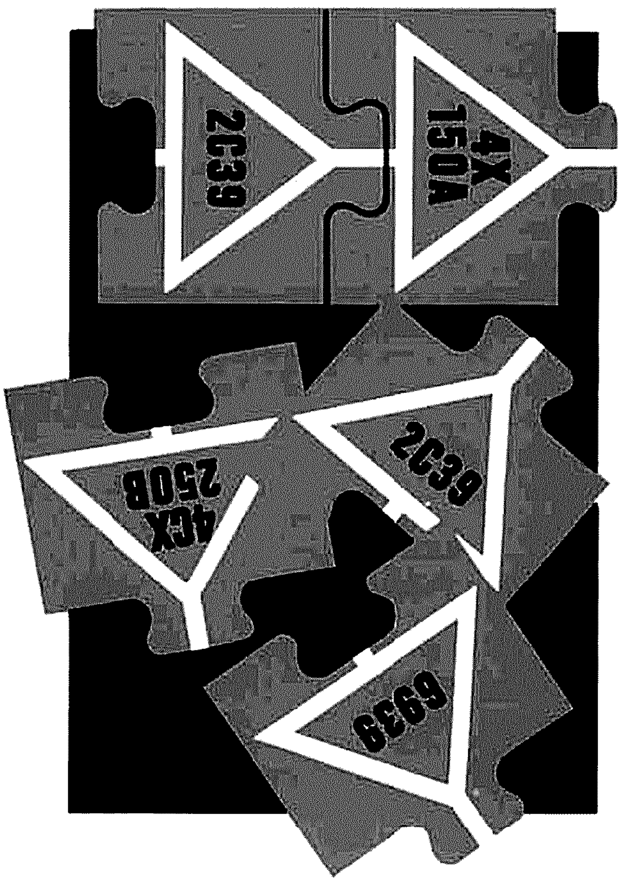
JUNE 1971

a practical approach to 432-MHz SSB



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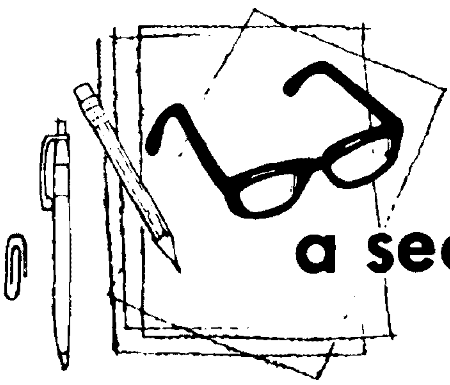


a second look

by jim fisk

A **whole new class** of broadband high-power rf amplifiers is about to hit the market; under study for some time, they use both vacuum-tube and semiconductor technology. The EBS – electron-bombarded semiconductor – amplifier consists of an electron-gun modulation system, semiconductor target and output coupling network, all within a glass or ceramic envelope. The semiconductor target is simply a pair of silicon diodes, each consisting of two metallic electrodes with a pn junction under the top contact.

Amplifier operation is based on a well-known fact: A modulated electron beam can control the current in a reverse-biased semiconductor junction. In the EBS system shown in **fig. 1** the electron beam is intensity modulated by an input signal on the grid. The high-power beam electrons that strike the silicon target create thousands of electron-hole pairs. Since there is a high bias voltage across the target, the free electrons are attracted to the far contact; the holes return quickly to the bombarded contact.

In the absence of the electron beam there is negligible current flow through the silicon. However, when illuminated by the electron beam, output current is proportional to beam current, and the device acts as a linear amplifier. The rf gain of the EBS amplifier is controlled by the current gain in the silicon target and the modulation sensitivity of the electron beam. Present laboratory amplifiers provide gains in excess of 40 dB with efficiencies over 65%.

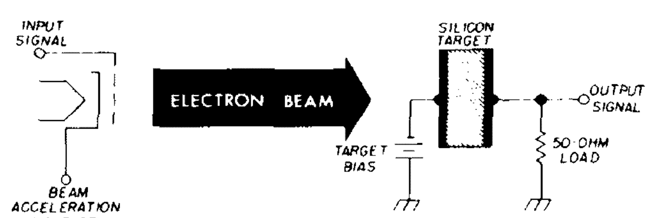


fig. 1. Basic EBS amplifier consists of modulated electron gun and semiconductor target.

The electron beam can also be deflection modulated as in conventional cathode-ray tubes. With a deflection-modulated EBS amplifier and two semiconductor targets, no beam current is intercepted when the beam is not modulated. With an input signal, the electron beam is deflected from one target to another; the positive portion of the sine wave is generated in one diode and the negative portion generated in the other. This is true class-B operation and provides high power-output capability.

The EBS amplifiers presently reaching the commercial market are limited to operation below 1500 MHz. However, engineers are working on new designs that should provide operation up to 5000 MHz. Although it will be some time before these devices find their way into the amateur market – and then probably by way of the surplus market – the EBS amplifier offers an alternate approach to the serious uhf amateur who wants to generate prodigious amounts of rf power on 1296 MHz and above.

Jim Fisk, W1DTY
editor

a practical approach to 432-MHz ssb

A survey of techniques and devices capable of generating ssb on 432 MHz

Fred Telewski, WA2FSQ, 460-40 Old Town Road, Port Jefferson Station, New York 11776

This article is the result of a one year study of techniques and devices applicable to 432 MHz ssb. I believe the techniques described here represent the best alternatives presently available. In each case I considered a number of approaches; the most efficient* are treated in detail, while the disadvantages of the others are discussed in an effort to acquaint you with some of the pitfalls. This by no means implies that the techniques not advocated are impossible; the non-recommended methods are feasible but at a much higher overall cost.

In an attempt to make this article as comprehensive as possible, all components were built in modular form. While this adds slight additional cost because of patch cables it permits the ultimate in versatility and allows verification tests of all circuit arrangements. Finally, all tests were made with calibrated laboratory test equipment. This provides a valid data base in areas which are normally troublesome to the amateur (i. e. spurious output and intermodulation distortion).

introduction

While operating on the 432-MHz band during the past several years I have noticed a definite lack of effective modulation on most signals. Although a large number of operators cling to cw (and its

*Efficient in terms of electrical efficiency, number of components and cost.

weak-signal effectiveness cannot be disputed) it disturbs me to see many contacts severely limited in their ability to exchange information when conditions permit the use of higher density modulation. Though amateurs are not hampered in their ability to generate power at 432 MHz they are apparently limited in their ability to modulate it efficiently.

It is the philosophy of this article to provide you with more than a cookbook which tells you where to cut, bend and drill. While much of the equipment discussed here may be duplicated it is my belief that few amateurs make identical copies. However, sufficient information is supplied to help you make intelligent decisions on the basis of economy and already existing equipment. Although the equipment discussed here was evaluated with laboratory-grade test equipment, the ssb units were aligned for maximum output power in an amateur manner, then evaluated, so the data is meaningful to the amateur without access to sophisticated test gear.

power considerations

It is helpful to acquaint yourself with the various devices that will function as amplifiers at 432 MHz. Linear uhf transistors are prohibitive because they are expensive and limited to low power.

Klystrons, amplitrons and cross-field amplifiers, while capable of high power, are not suitable due to cost and complexity. This leaves us with the family of power-grid tubes. **Table 1** shows popular vacuum tubes and approximate gains which may be obtained as 432-MHz linear amplifiers. To make this information pertinent to the amateur both new and used tubes were tested.

The third-order intermodulation distortion (IMD) values* for the tubes I tested fall in the 20- to 30-dB range. While this may sound low in comparison to 20-meter standards, an objective analysis is necessary to determine what ill effects if any this will have on 432-MHz operation. The major objections to IMD are its generation of signals outside the communications passband and wasted power. The former is a particular concern on the high-frequency bands because of

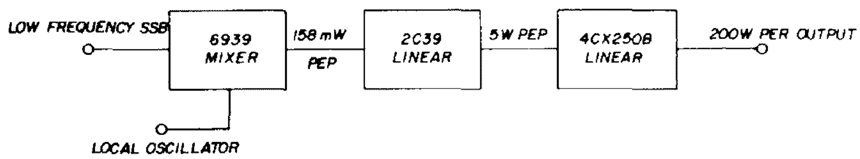


fig. 1. Proposed system for running 500 watts PEP input on 432 MHz. Efficiency of final stage is assumed to be 40%.

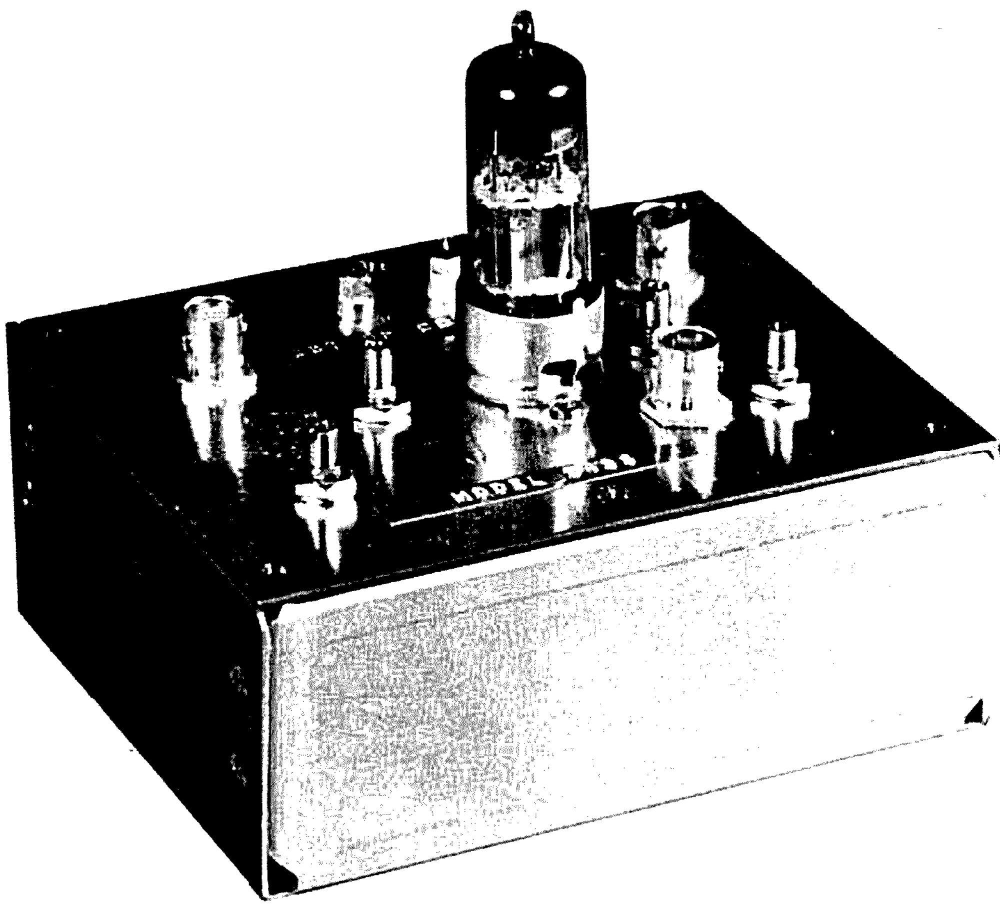
*The intermodulation distortion figures given here were measured in accordance with standard commercial practice: in reference to one of the two test tones. Amateurs normally measure intermodulation distortion in respect to peak envelope power; this yields IM figures 6 dB better than commercial practice. To compare these IM figures with other amateur radio articles you must increase the IM figure by 6 dB. In this case, 20 dB becomes 26 dB.

table 1. Typical linear performance of several readily available power-grid tubes.

type	gain (dB)	power output (PEP watts)	3rd order IM (dB)
6939	15	2.6	25
2C39, 2C39A	12.5-15	30	25-30
2C39B, 3CX100A5, etc.	14-16	40	25-30
4X150A (new)	15	80	20-25
(used)	14	80	20-25
4CX250B (new)	17	200	20-25
(used)	16	200	20-25

high signal densities. However, it is hardly worth mentioning at uhf. Wasted power seems to be of little consequence since an IMD ratio of 20 dB results in a 1% power loss. Most amateurs lose that adjusting his station for maximum power output.

Considering a medium power system, 500 W PEP input with final-amplifier efficiency of 40%, the output power



432-MHz ssb mixer using a 6939.

much power in the first few feet of transmission line. Consequently, all designs presented here are based on achieving less than 20-dB IMD at the output, and assume that exciter IMD is 30 dB down.

IMD performance is given constant attention throughout this article. Measured values of IMD are presented for various levels of operation. It should be noted that all IMD tests were made with the equipment optimized for maximum power output, *not* minimum IMD. Therefore, the measured values are in line with what an amateur might expect when should be in the neighborhood of 200 W PEP or +53 dBm. A 4CX250B is capable of 16-dB gain, so its drive requirement is +37 dBm (5W PEP). A good 2C39 linear will drop the drive requirement to +22 dBm (158 mW PEP). At this point assume a mixer which will produce between 0.1 and 1 W PEP output.

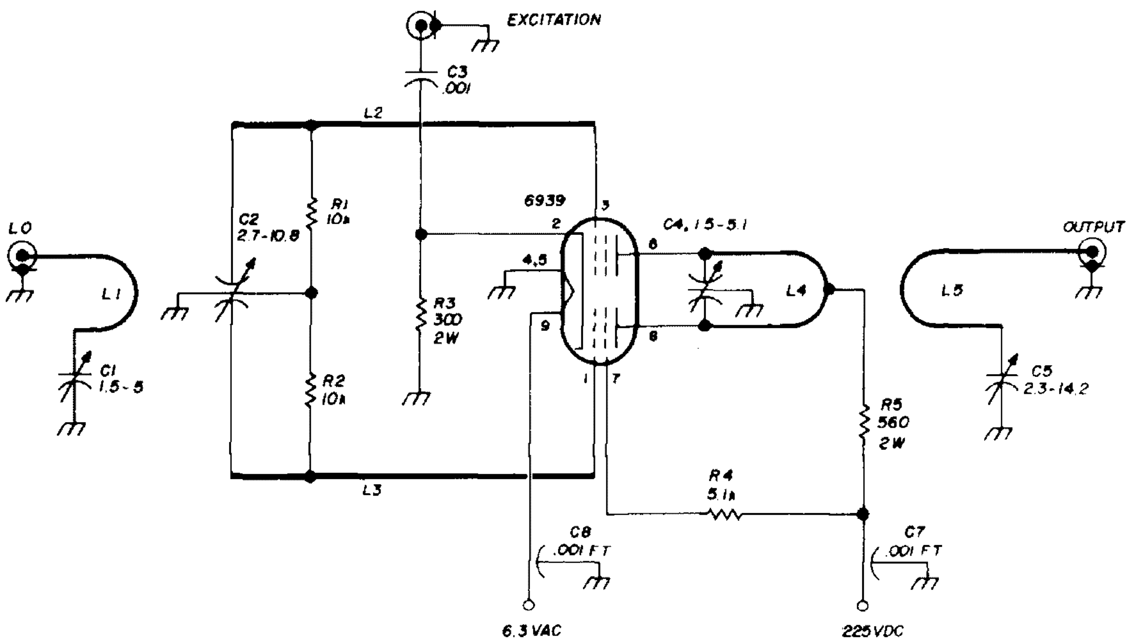
Fig. 1 shows the proposed system. A power train of three stages will take low-frequency ssb and put it on 432 MHz at the 500-watt level. At this point, it appears to be no more difficult to put 500 watts on 3/4 meters than it is on 20 meters.

the mixer

The search for a suitable mixer covered a great deal of territory. It must be efficient, be stable, exhibit low IMD, be easily reproduced and, perhaps most important, have low cost. Although I considered parametric upconverters, and did make one work in the laboratory, the parametric device could not be deemed a reproducible unit for the amateur without access to sophisticated test equipment (the unit also required 15 watts of pump power at 382 MHz). The second consideration was in the area of transistor mixers. Those transistor stages which were economical operated at low power levels (less than -10 dBm) and required an unreasonable number of linear gain stages to achieve respectable output power. Those stages which could handle higher power were cost prohibitive. Various diode mixer configurations were also examined but their limitations fall in the first transistor category. This narrowed the mixer field to our old friend the vacuum tube.

The Amperex 6939 seemed to be a good choice since it has already been successfully used as a high-frequency mixer.[1] However, it seems that most of the published designs are limited by their high local-oscillator power requirements. The circuit shown here (fig. 2) is somewhat similar to those previously published but may be driven with approximately 100 mW of local oscillator power instead of the usual watt or two.

Many of the problems encountered with vacuum tubes are in the area of coupling power into high-capacitance grid structures. You can get an idea of the efficiency of any matching network by measuring its reflection coefficient when



C1	1.5-5 pF (Johnson 160-102)
C2	2.7-10.8 pF (Johnson 160-211)
C4	1.5-5.1 pF (Johnson 160-205)
C5	2.3-14.2 pF (Johnson 160-107)

fig. 2. 432-MHz mixer stage provides high efficiency. Stage requires approximately 100 mW local-oscillator power. Inductor construction is shown in fig. 3.

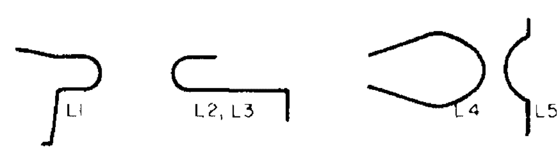
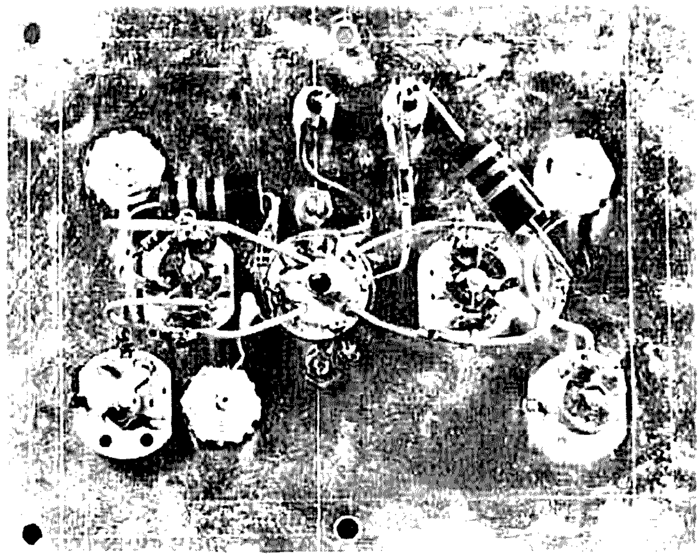


fig. 3. Coils for 6939 mixer (fig. 2) and linear-amplifier (fig. 10).*

*Full-size templates of fig. 3, fig. 8 and fig. 18 are available from **ham radio** for 25c.

table 2. 6939 mixer operating parameters (plate dissipation, 6 watts).

Static		
plate voltage	(E_b)	225 Vdc
plate current	(I_b)	23 mA
Driven with local oscillator		
plate voltage	(E_b)	225 Vdc
plate current	(I_b)	28 mA
screen voltage	(E_{c_2})	210 Vdc
screen current	(I_{c_2})	6 mA
cathode voltage	(E_k)	9 Vdc
grid current	(E_{c_1})	0 Vdc



Construction of 6939 mixer stage.

adjusted for maximum output power. It is a well known fact of network theory that any lossless, reciprocal, passiye, two-port network will have its output reflection coefficient equal in magnitude to its input reflection coefficient. Since our networks are certainly passive, reciprocal, and hopefully, *not too lossy,* this is a good beginning for optimizing any grid circuit.

An idea of grid-circuit loss may be obtained by shorting the grid to ground (with B+ and bias off) and noting the input vswr; it should be greater than 10:1 if circuit losses are low. With the tube operating normally the input vswr should be less than 2:1 for a reasonable grid circuit. In general, the problem of coupling to a high capacitance grid structure may be easily dealt with through the use of half-wave lines.[2] Quarter-wave circuits are usually ineffective since tube capacitance severely forshortens the line and it is difficult to place a tuning capacitor sufficiently close to the grid for proper operation.

The half-wave grid lines used in this mixer provide a vswr of 1.3:1 or less after one or two passes of the tuning wand. The rest of the circuit is quite conventional. The tube is operated within the normal ratings specified by the manufacturer consistent with reliability and economy. Table 2 lists the operating parameters of the 6939 mixer tube.

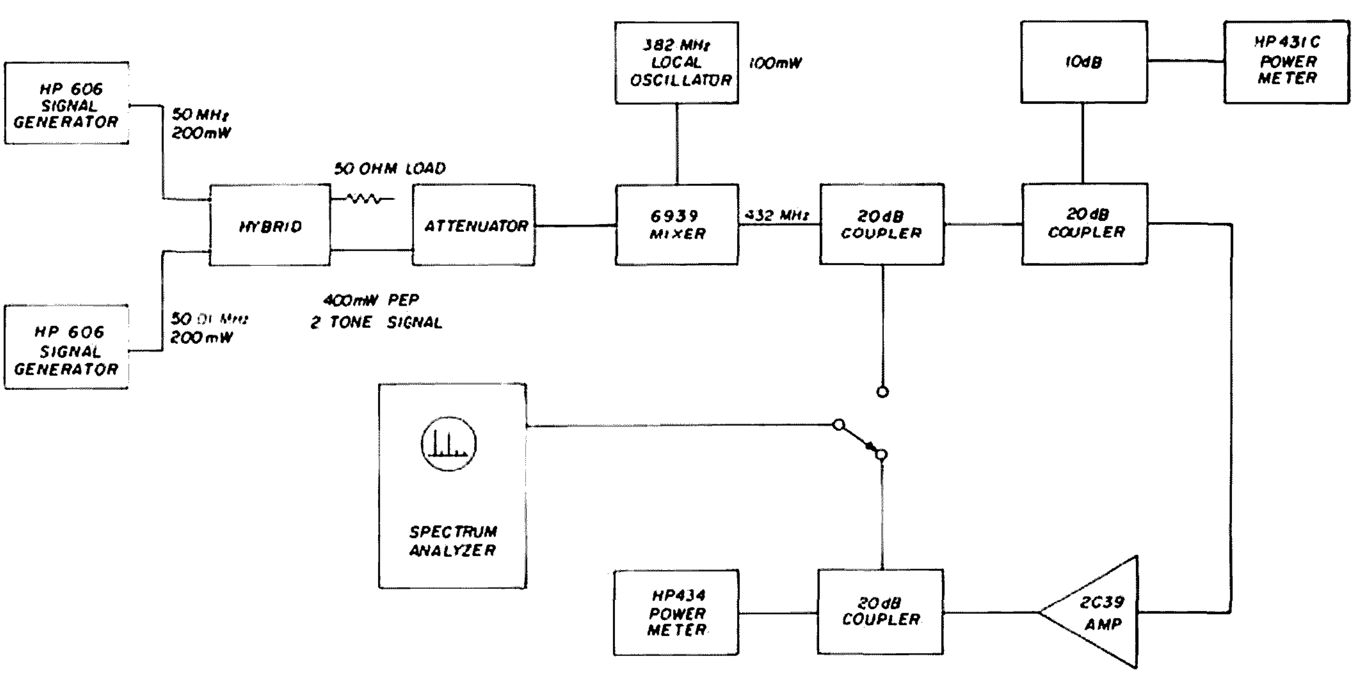


fig. 4. Setup for spurious output and intermodulation distortion tests.

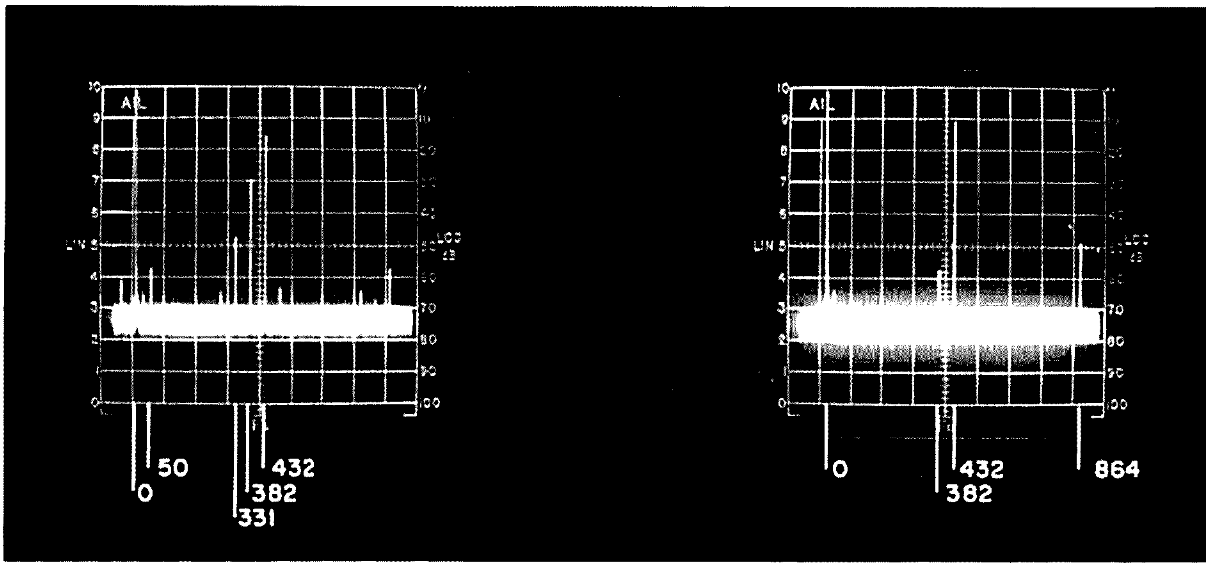


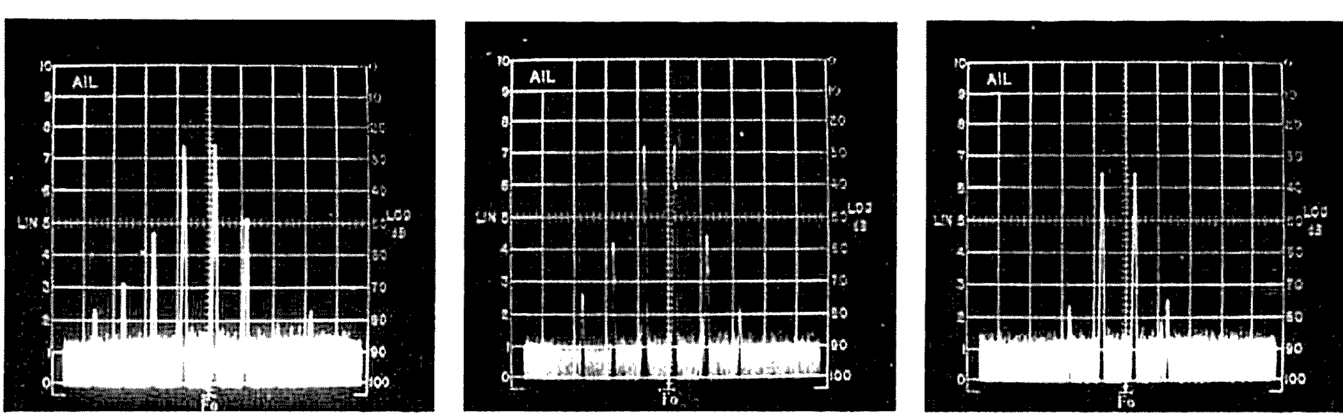
fig. 5. Spectrum analyzer display of 6939 mixer and 2C39 amplifier; horizontal, 100 MHz per division. 6939 mixer output, left, with 100 mW local-oscillator power, 400 mW PEP drive and 240 mW PEP output on 432 MHz. 2C39 output, right, with 10 watt PEP output on 432 MHz.

mixer performance

A few words concerning spurious mixer outputs are in order. The diagram in **fig. 4** describes the spurious and IMD evaluation tests for 50 MHz excitation. Directly at the mixer output the local oscillator is suppressed 15 dB and the image is down 32 dB (see **fig. 4**). At the output of the first gain stage local oscillator feedthrough is suppressed 45 dB and the image is down more than 60 dB.

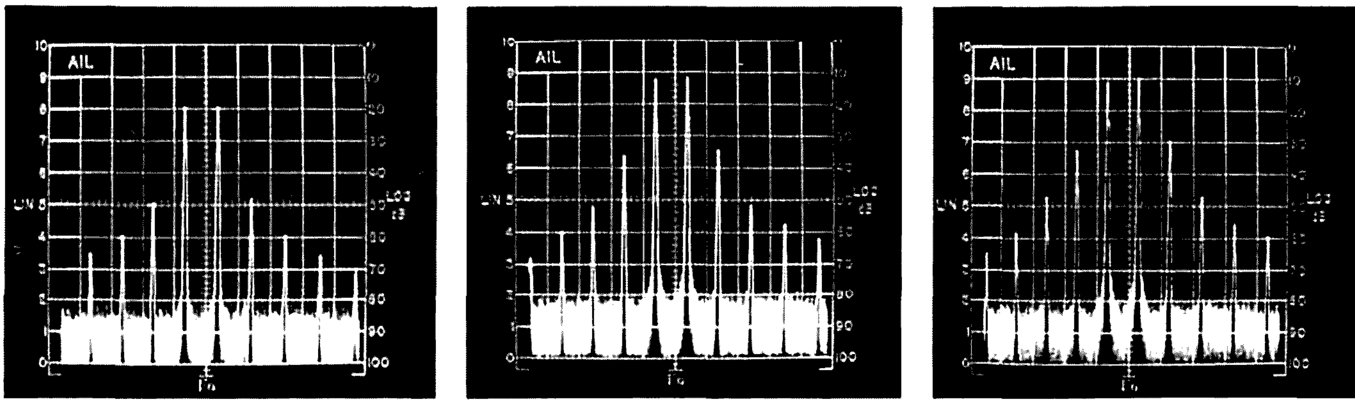
If 28 MHz excitation were chosen, at the output of the first gain stage the local oscillator would be suppressed 25 dB, and the image would be 50 dB down. Considering the power level and typical off-resonance antenna efficiencies these numbers are adequate. With excitation below 28 MHz an interdigital filter such as the one described by W2CQH[3] should be used between the mixer and the first linear gain stage. With an interdigital filter in the system an excitation frequency as low as 9 MHz will provide spurious suppression superior to that obtainable with 28 MHz excitation without the filter.

The IMD performance of the 6939 mixer at various power levels is sum-



local oscillator	100 mW	local oscillator	100 mW	local oscillator	100 mW
two-tone drive	40 mW PEP	two-tone drive	240 mW PEP	two-tone drive	400 mW PEP
432-MHz output	28 mW PEP	432-MHz output	160 mW PEP	432-MHz output	240 mW PEP
3rd-order IMD	-40 dB	3rd-order IMD	-29 dB	3rd-order IMD	-23 dB

fig. 6. 6939 mixer intermodulation distortion at various power levels.



432-MHz drive	28 mW PEP	432-MHz drive	160 mW PEP	432-MHz drive	240 mW PEP
432-MHz output	1 W PEP	432-MHz output	6 W PEP	432-MHz output	10 W PEP
3rd-order IMD	-30 dB	3rd-order IMD	-24 dB	3rd-order IMD	-20 dB

fig. 7. 2C39 amplifier intermodulation distortion tests. These tests were run simultaneously with the 6939 mixer test shown in fig. 6.

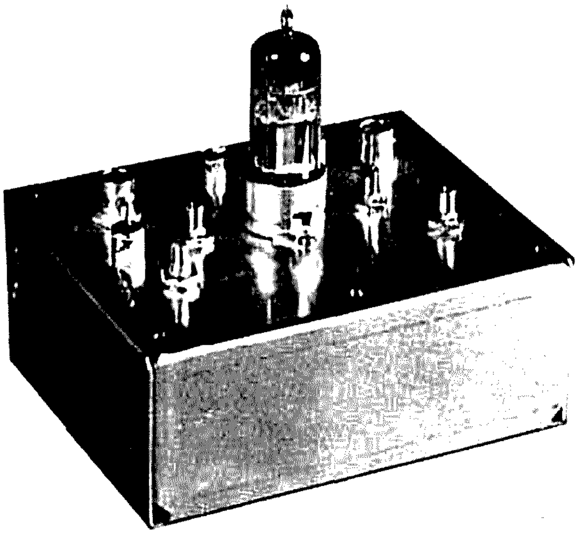
marized in fig. 6. The 6939 mixer, when operated according to table 2, with 240 mW PEP 50-MHz drive, is capable of 150 mW PEP output; IMD under these conditions is 30-dB down.

A chassis layout for the 6939 mixer is shown in fig. 8. If you use all new components in the construction of the mixer, total cost will be less than $20. The 6939 tube, the most expensive part, lists at $12.

linear amplifiers

Depending on the amount of power you want to put on 432 MHz, the linear amplifier can take any of the forms

6939 linear amplifier for 432 MHz.



shown in fig. 9. This chart shows various vacuum-tube combinations for power level from 5 W PEP to 2 kW PEP.

The 6939 linear amplifier stage in fig. 10 is capable of 6 watts PEP input with approximately 2.6 W PEP output. This represents a stage efficiency of 43%. This design uses essentially the same circuit as the mixer except that the cathode is directly grounded, an external bias port is provided for the grid, and the screen is fed from a regulated source (to improve linearity). Mechanically, the 6939 mixer and amplifier layouts differ only by three holes, so both chassis plates may be drilled simultaneously. Table 3 lists the typical operating parameters of the 6939 amplifier.

the 2C39

I compiled a good deal of information about the operation of the 2C39 family. Older tubes such as 2C39s and 2C39As, and used tubes of newer varieties, offer an interesting bonus: they may be operated as zero-bias triodes. New 2C39Bs, 3CX100A5s, etc., require some bias to achieve reasonable plate currents. While best linearity and maximum output power occur when the tubes are idled at approximately 60% of their rated dissipation (60 W), they may be idled as low as 15 watts with reduced power output.

Under zero-bias conditions this dissipation range typically corresponds to a

plate voltage range of 350 to 700 Vdc and output powers from 4 to 30 watts PEP. The gain of the tube varies about 1 dB between these two voltage extremes with the higher gain at the higher operating voltage. **Table 1** and **fig. 11** detail the typical properties of the 2C39 family. The dc operating parameters of the 2C39 stage used earlier (in the IMD and spurious products evaluation) were, E_b = 400 Vdc, I_b = 60 mA, plate dissipation = 24 watts. Spurious and IMD characteristics of this 2C39 amplifier are summarized in **fig. 7**.

Since the 2C39 amplifier I used to compile the 2C39 data uses a surplus cavity, no actual homebrew amplifier is described in this article. However, several

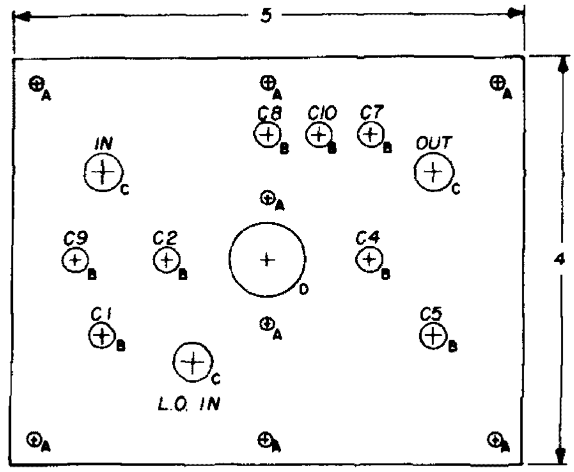


fig. 8. Layout template for 6939 mixer (fig. 2) and linear amplifier (fig. 10). For the mixer circuit omit holes for C9 and C10. For the linear amplifier omit the local-oscillator input. Hole sizes; A, 0.140"; B, 0.250"; C, 0.375"; D, 0.750".

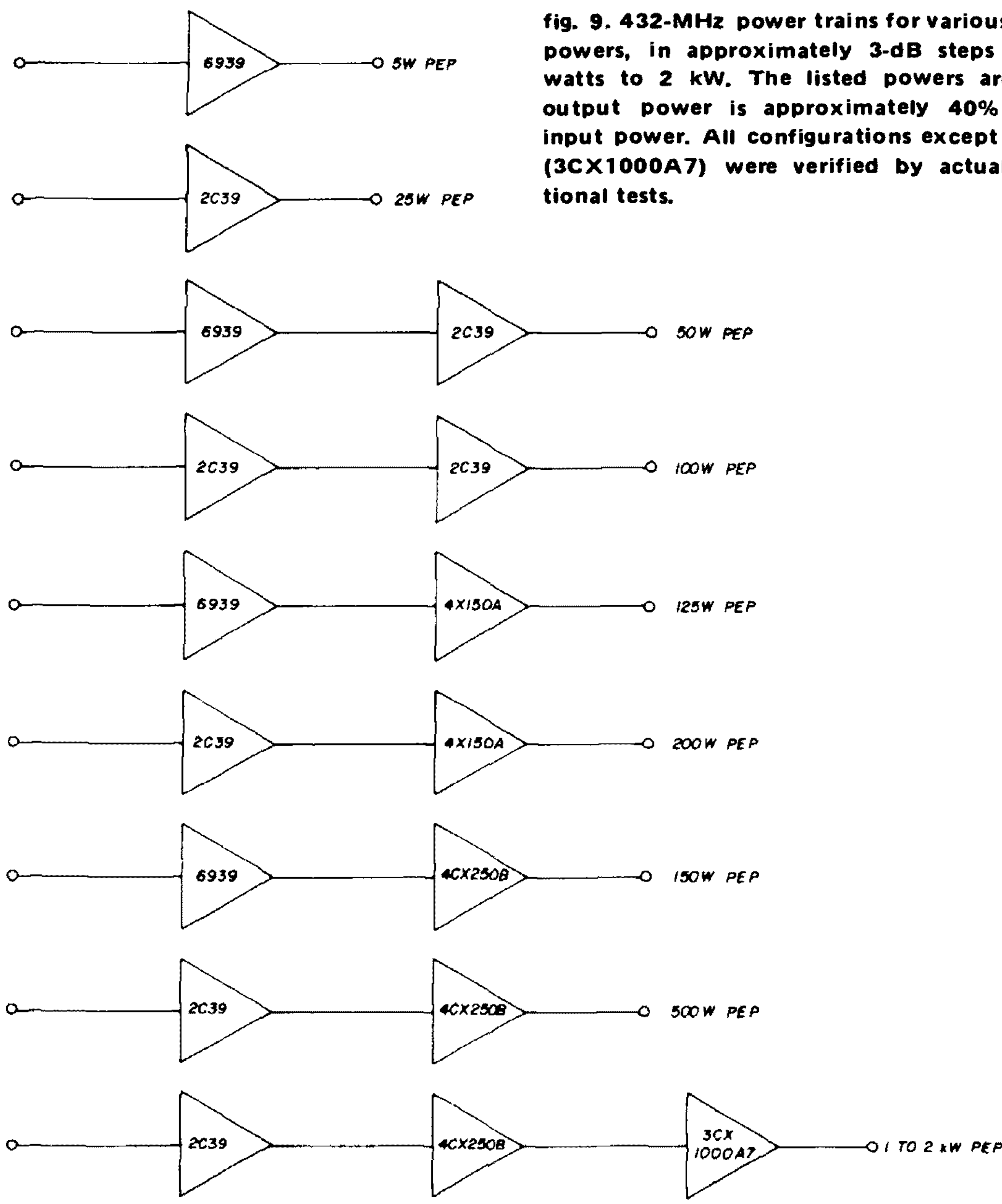


fig. 9. 432-MHz power trains for various output powers, in approximately 3-dB steps from 5 watts to 2 kW. The listed powers are input; output power is approximately 40% of the input power. All configurations except the last (3CX1000A7) were verified by actual operational tests.

table 3. 6939 linear amplifier operating parameters. (Power dissipation approximately 6 watts.)

plate voltage	(E_b)	225 Vdc	
plate current	(I_b)	28 mA 32 mA	(static) (dynamic)
grid bias	(E_{c_1})	3.5 Vdc	
screen current	(I_{c_2})	6 mA 15 mA	(static) (dynamic)

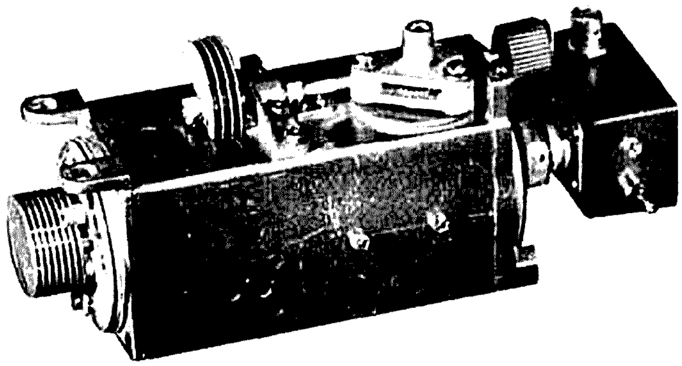
compact 2C39 designs appear in the W6AJF vhf handbook.[5] These circuits, when used with half-wave cathode-drive circuits, function as well as the surplus cavity I used in my tests.

higher power

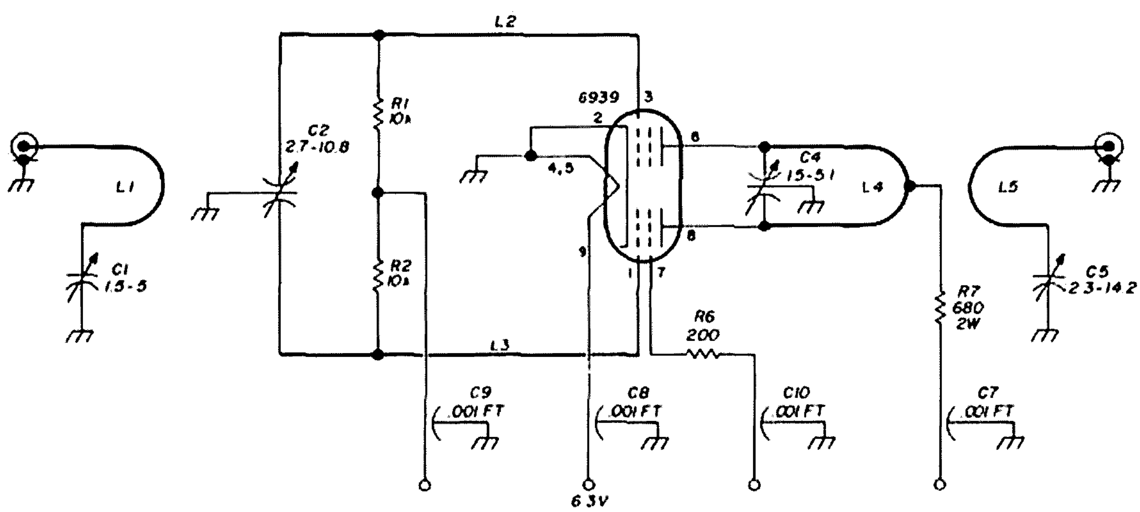
The 4X150A and 4CX250B pave the way to higher power on 432 MHz. Both tubes, when properly used, exhibit relatively high power gains (see table 1). The 4X150A is capable of approximately 200 PEP input at 432 MHz while the 4CX250B is capable of 500 W PEP input.

In keeping with the philosophy of this article, good design practices will be discussed first. This will allow you to examine your present 432-MHz gear and make the necessary modifications.

The first consideration is that of driving the tube. To start with, avoid using a large grid compartment as this may lead to many problems. Much of the rf drive may find its way into resistors, filament transformers and other dissipative structures within the compartment. Another problem which tends to plague large grid compartments is that of multimoding. When the size of the grid compartment happens to coincide with certain preferred dimensions the entire compartment may act as a resonator. Usually the



Surplus 2C39 cavity used in the 432-MHz tests.



C1	1.5-5 pF (Johnson 160-102)	C4	1.5-5.1 pF (Johnson 160-205)
C2	2.7-10.8 pF (Johnson 160-211)	C5	2.3-14.2 pF (Johnson 160-107)

fig. 10. 6939 linear amplifier operates at 43% efficiency. Inductors for this stage as shown in fig. 3; layout is in fig. 8. Operational data for this stage is listed in table 3.

result is multiple tuning peaks and erratic grid-network behavior. Incidentally, multimoding is also a problem in plate tank compartments which are too large. It is a good rule of thumb to keep two dimensions of the box well below one-half wave at the operating frequency to eliminate

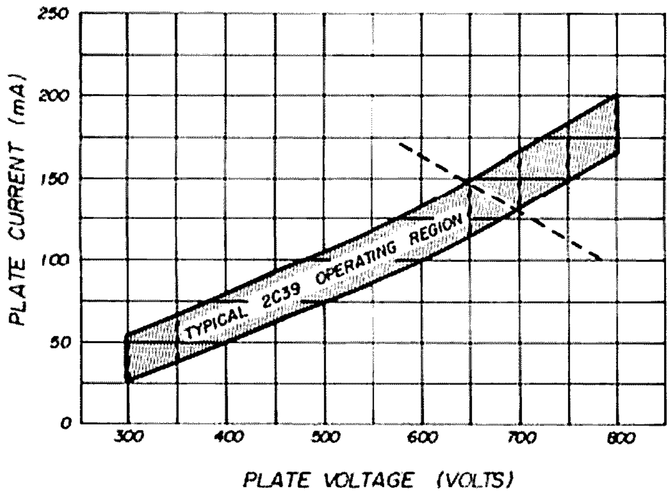


fig. 11. Zero-bias operation of tubes in the 2C39 family. To prevent excessive dissipation, bias should be used if the tube operates to the right of the dashed line.

multimoding (at 432 MHz this corresponds to approximately 13.5 inches).

grid circuit

Eimac engineers, when designing the SK610 socket, were kind enough to provide a means of fastening 1-5/8-inch tubing to the grid side of the socket, thereby allowing you to make a grid cavity. Since 1-1/2 inch copper plumbing pipe makes a snug fit to the base of the socket it's easy to make an efficient grid cavity.

You must now decide whether to use a 1/4- or 1/2-wave grid cavity. (Eimac Application Bulletin no. 14 gives many fine examples of 1/2-wave grid structures which are efficient and simple to construct.) In the amplifier discussed here a somewhat different approach to a 1/4-wave structure is taken.

The input reactance of the 4X150A, 4CX250B series of tubes is in the vicinity of 25 ohms capacitive. The length of the grid circuit in electrical degrees may be calculated from the equation:

$$\theta = \tan^{-1} \frac{X_C}{Z_O} \qquad (1)$$

where X_C is the capacitive reactance (25 ohms) and Z_O is the characteristic impedance of the cavity. For the coaxial arrangement used here Z_O is given by the equation:

$$Z_O = 138 \log \frac{b}{a} \qquad (2)$$

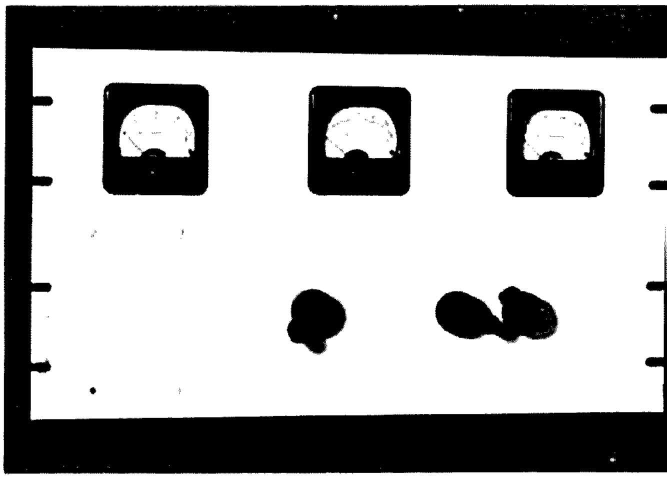
where b is the inside diameter of the outer conductor and a is the outside diameter of the inner conductor. K7UNL provided design data for other types of resonant lines in a recent article in *ham radio.*[6]

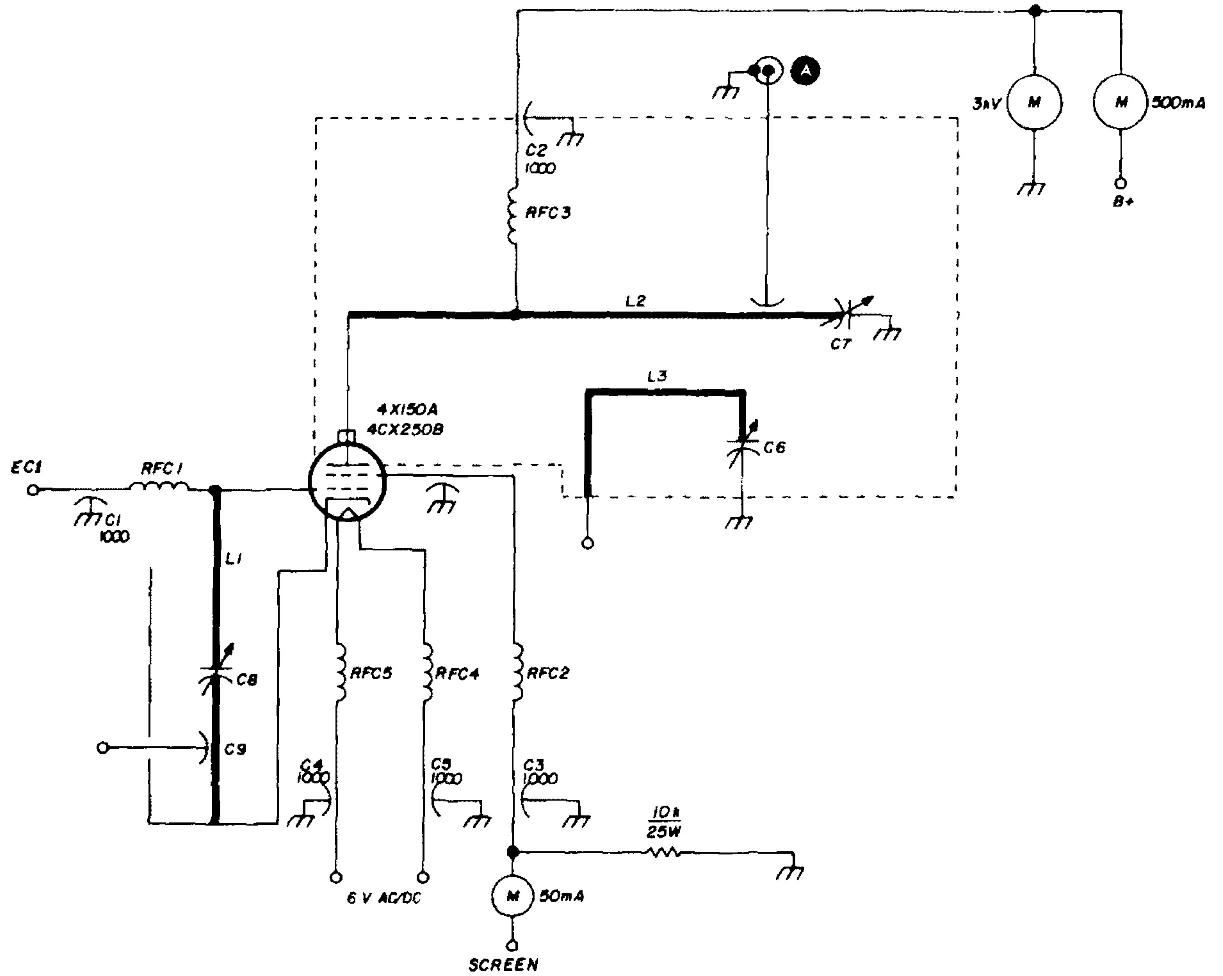
Choosing an inner conductor diameter of 3/4 inch and the outer conductor diameter of 1-1/2 inch, Z_O is approximately 41 ohms. Running this through eq. 1 yields 31.4°. Proportion will provide the physical length of the line:

$$\frac{31.4^\circ}{360^\circ} = \frac{l}{\lambda} \qquad (3)$$

where λ = 69 cm. (Equivalent wavelength for 432 MHz is 69 cm.) Therefore l = 6 cm or 2.3 inches. Unfortunately, this is a bit short and as yet there is no way of tuning the grid circuit. This verifies previous comments about quarter-wave lines being severely forshortened by high capacitance grid structures.

Front-panel layout of the 432-MHz power amplifier. Circuit is shown in fig. 12.





C6	loading capacitor. 0.4-8 pF piston (JMC1802 may be used) homemade capacitor used in amplifier in photos	C9	1-1/4" length 1/4" brass rod soldered to BNC connector, braced with 1/4" Teflon rod (see fig. 13)
C7	air disc capacitor. 1-3/4" diameter, 1/8" thick brass disc soldered to 3½" length of 3/4" diameter threaded bronze pipe. Pipe is mounted through threaded bronze flange (available at plumbing supply) (see fig. 13)	L1	grid cavity (see fig. 13)
		L2	copper plate line (see fig. 15)
		L3	copper output strap (see fig. 15)
		RFC1	8 turns no. 22 enameled wire closewound on 1/8" form
C8	1" diameter, 1/8" thick brass disc capacitor (see fig. 15)	RFC2,3,4,5	8 turns no. 20 Teflon-insulated solid wire closewound on 3/8" form

fig. 12. 432-MHz linear amplifier stage using a 4X150A or 4CX250B. Alternate output coupling at A. Screen bypass capacitor is part of tube socket.

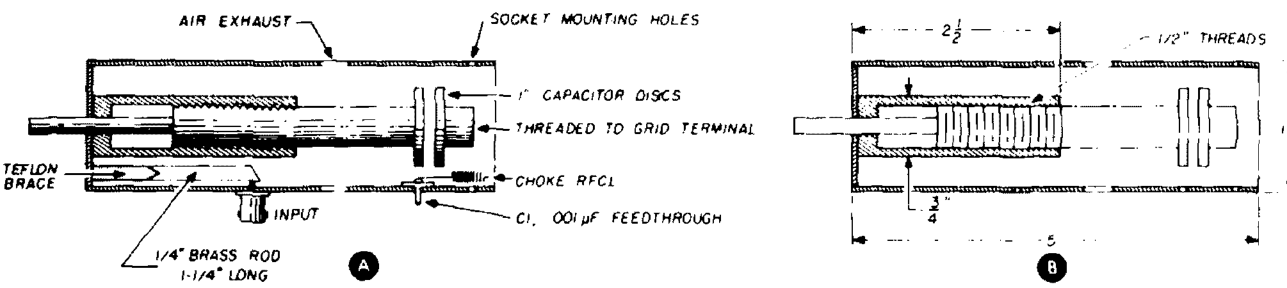


fig. 13. Tuned grid circuit for the highpower 432-MHz linear amplifier.

table 4. 4CX250B linear operating parameters. For the 4X150A, maintain 6-volt filament voltage and adhere to manufacturer's reduced ratings for uhf service.

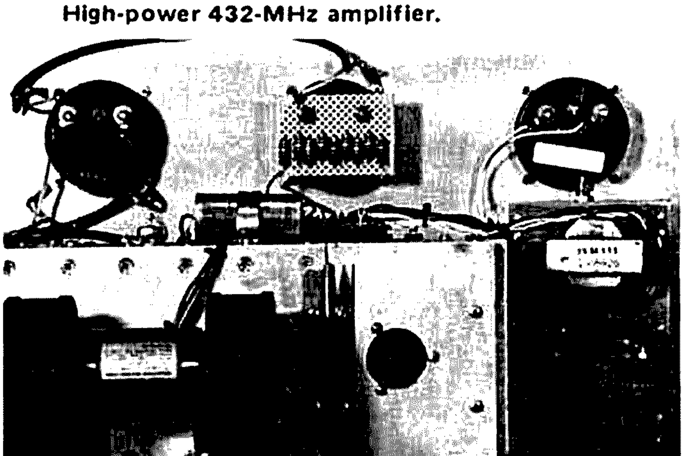
power input		375 W	500 W
plate voltage	(E_b)	1500 Vdc	2000 Vdc
screen voltage	(E_{c2})	350 Vdc	350 Vdc
grid voltage	(E_{c1})	-55 Vdc	-55 Vdc
plate current	(I_b)	100 mA 250 mA	100 mA (static) 250 mA (dynamic)
screen current	(I_{c2})	8 mA	5 mA

At this point, you would probably be tempted to install a variable capacitor across the high-voltage end of the line to tune it. This is undesirable as it would further shrink the already forshortened line. The solution to the tuning problem is a series variable capacitor between the grid and the line. Two parallel discs 2.5 cm (1-inch) diameter, spaced 1/2 cm apart exhibit approximately 9 pF capacitance. Using the equation for series capacitors

$$\frac{1}{C_t} = \frac{1}{C1} + \frac{1}{C2} + \ldots \frac{1}{C_n} \qquad (4)$$

The total capacitive reactance across the line is approximately 60 ohms. With 60 ohms as a basis, the recomputed line length (using eq. 1 and 3) is 4 1/4 inches. This is a reasonable length, and the grid resonator is now tunable.

High-power 432-MHz amplifier.



The overall length of the grid line is increased to 5 inches to account for the 1 cm spacing between the plates, the thickness of the plates (approximately 1/4 inch) and any non-computed second order effects (see fig. 13). Remember, if the line is slightly longer than necessary it may easily be shortened; however, if the line is too short to begin with you will have trouble trying to stretch it!

Another interesting problem associated with these tubes is the matter of forced-air cooling. In linear service it is advisable to use maximum recommended plate voltage to obtain maximum linear power output. Under these conditions the tube runs hot, and cooling it requires large amounts of air – provided by a moderately large fan or a smaller one operating at high speeds. Large fans are expensive

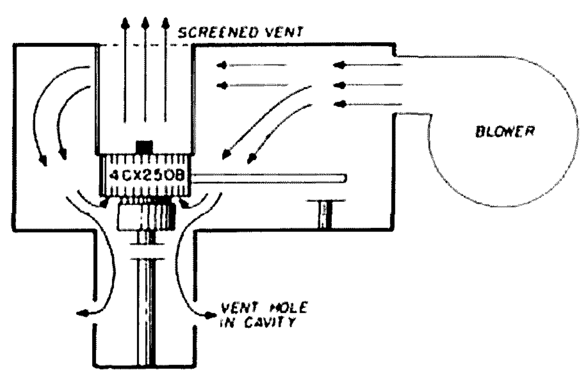


fig. 14. Air flow in the 432-MHz amplifier.

and bulky while small high speed ones are noisy. However, if you pressurize the plate compartment, disregard the conventional chimney and provide an exhaust passage from the anode radiator to the outside world (see fig. 14), the back pressure on the fan is considerably reduced. Under these conditions, the tube operates reasonably with a 3-inch 3600-rpm squirrel-cage fan at a pressure drop of 1.30 inch. This provides an added bonus in that the plate-tank components are air cooled; therefore the amplifier is free of resonance drift associated with the thermal expansion of the tank circuit.

The plate tank enclosure is made from 1/8-inch 2024-T3 aluminum stock with 1/2 x 1/2 x 1/8-inch aluminum angle at

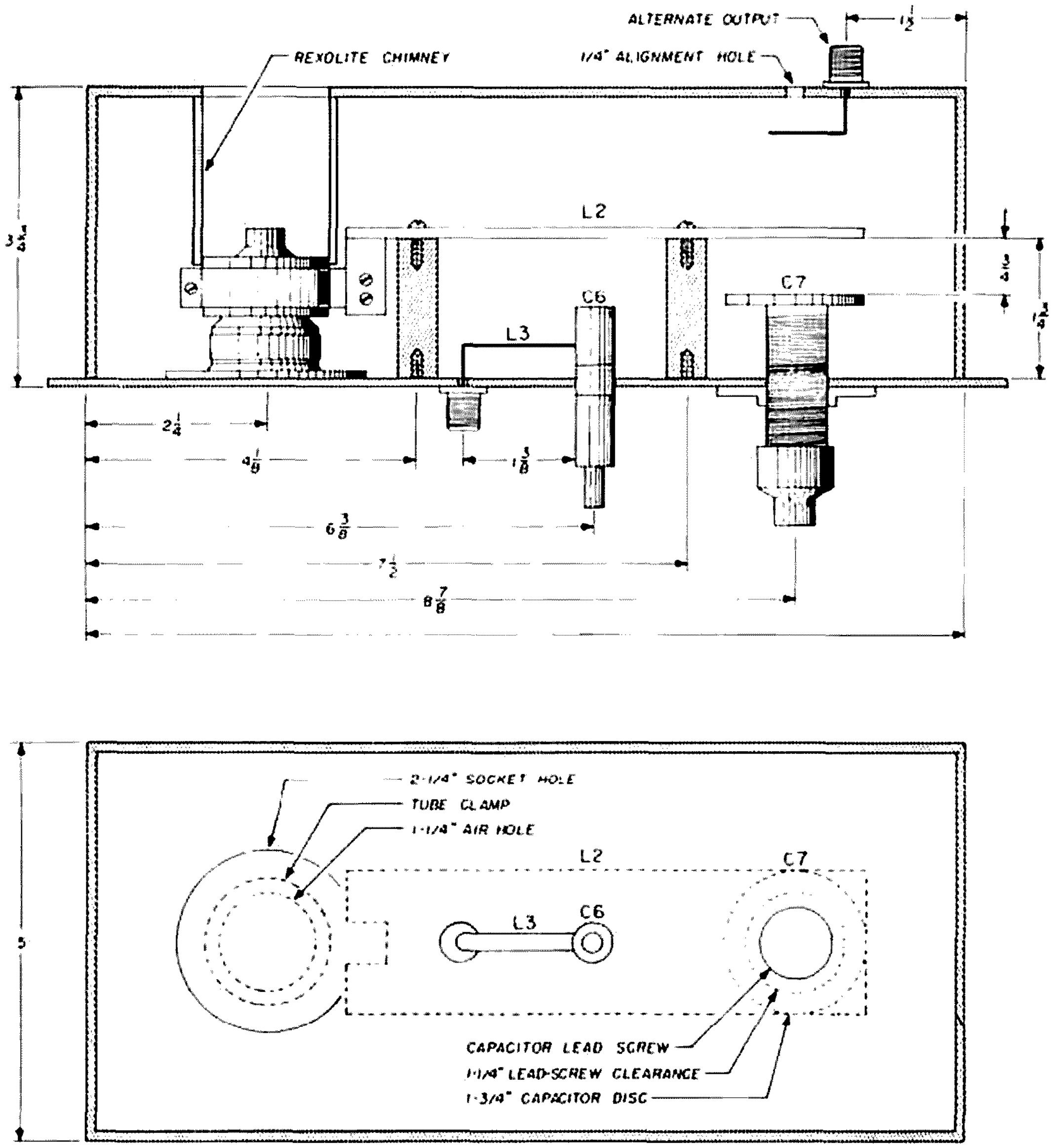
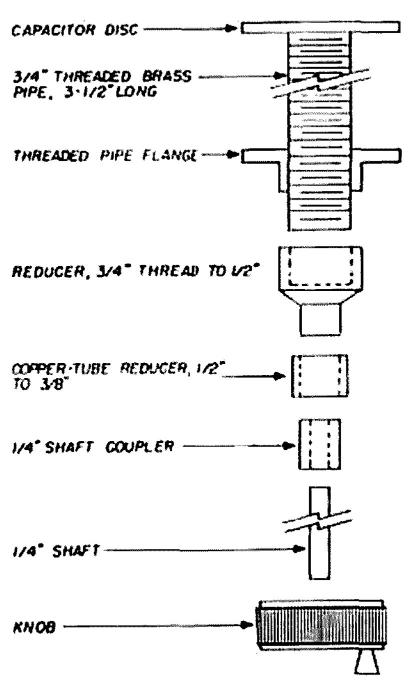


fig. 15. Layout details of the high-power 432-MHz power amplifier (fig. 12). Construction of capacitor C7 is shown at right.



the corners. The plate tank circuit is a conventional half-wave line. Output coupling may be accomplished with equal efficiency through inductive or capacitive probes as shown in fig. 12. Operating parameters are given in table 4.

filament voltage

Normally you reduce filament voltage at uhf to compensate for back bombardment of the cathode. Fortunately, back bombardment is minimized under the condition of linear operation. The usual criteria (i. e. minimum grid bias, high screen voltage and low grid drive) which minimize back bombardment are en-

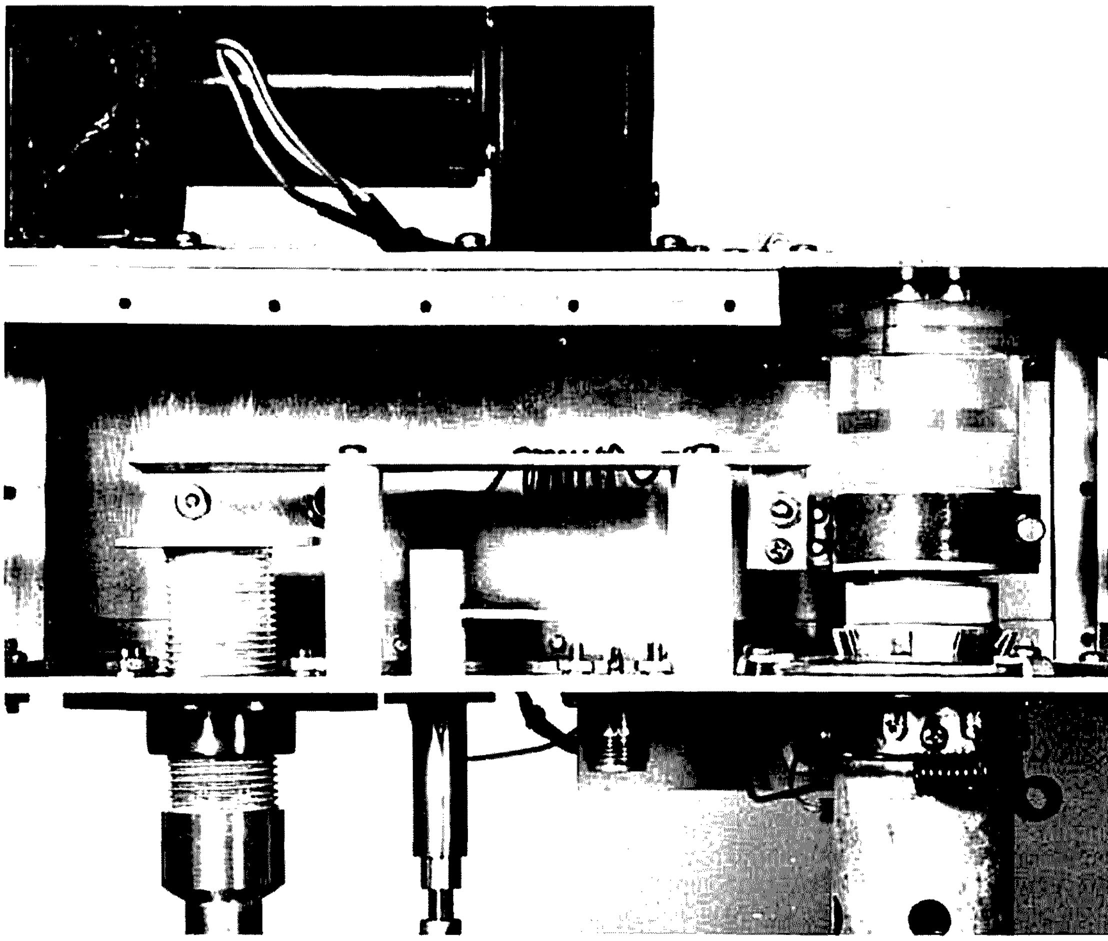


Plate compartment of the 432-MHz power amplifier.

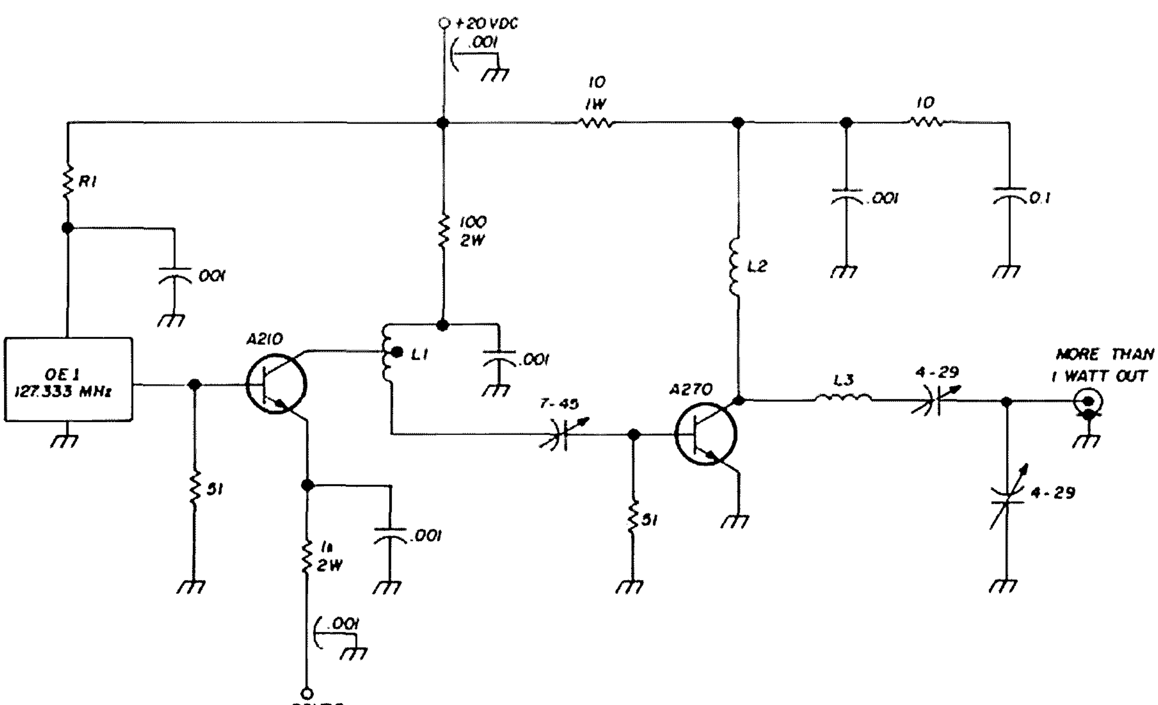


fig. 16. 127-MHz local oscillator uses International Crystal OE1 crystal oscillator with two-stage amplifier. Value of R1 is selected to give proper operating voltage for the OE1.

L1	6 turns no. 18, spaced one diameter, on 3/8" form
L2	1 μH choke
L3	3 turns no. 16 enameled close-wound on 1/2" form

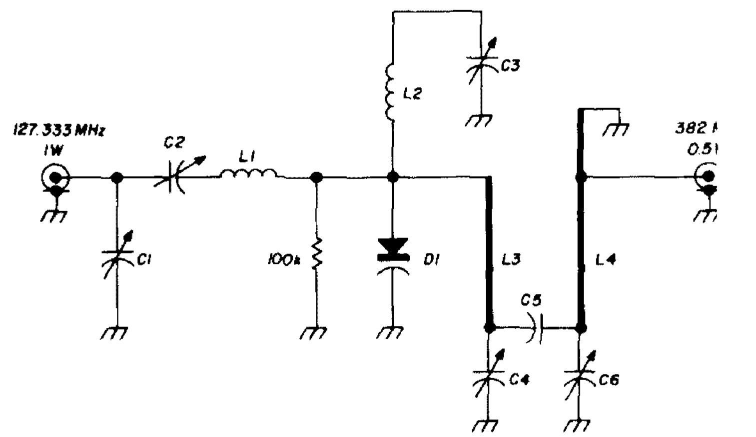
forced in linear service. At the same time, in linear ssb service, the duty cycle of the amplifier is significantly less than 100% thereby further reducing back heating.

Consequently, the filament voltage is maintained at the normal 6V. Should you desire to operate this amplifier class C, you must remember to reduce the filament voltage accordingly. However, there is no practical reason for using class C since you already have the most effective voice system available. If you wish to operate cw the linear will not deliver sufficiently less power than its class C counterpart to make bias and drive-level changes worth-while.

local oscillator chain

The solid-state local oscillator chain described here uses an International Crystal OE1 overtone oscillator at 127.333 MHz followed by an amplifier (fig. 16) and tripler (fig. 17) to produce power at 382 MHz, which when hetrodyned with a 50-MHz ssb exciter provides 432-MHz ssb. Other exciter frequencies (anything from 9 to greater than 50 MHz) may be used by altering the local-oscillator frequency. The circuit in fig. 16 uses two Amperex transistors to amplify the 1 mW OE1 output to the 1-watt level.

The oscillator amplifier was put together on a 4 x 6-inch piece of double-sided copper-clad board and mounted upside down in a 2 x 4 x 6-inch aluminum chassis. The 1 W output drives a varactor tripler which provides the 100 mW required for the mixer. A varactor tripler is used with the contingency that many amateurs presently on 432 probably have a varactor tripler which could be tuned down to the local-oscillator frequency, thereby saving a component. An equally adaptable vacuum-tube local-oscillator chain appeared in an article by K6JC.[1]



C1,C2	2.7-19.6 pF (Johnson 160-110)
C3,C4,C6	0.9-7.0 pF piston (JFD VC1G)
C5	gimmick. No. 20 wire, twisted, 1" long
L1	7 turns no. 18, spaced 1 wire diameter, 3/8" form
L2	2 turns no. 18, spaced 1 wire diameter, 3/8" form
L3,L4	hairpin made from 2" length of no. 18 wire. Hairpin 7/16" wide, 7/8" long
D1	Amperex H4A or equivalent

fig. 17. Varactor tripler is used with 127-MHz local oscillator (fig. 16) to provide injection at 382 MHz. Layout for this circuit is shown in fig. 18.

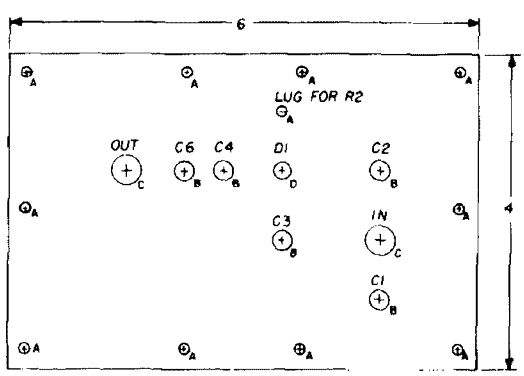


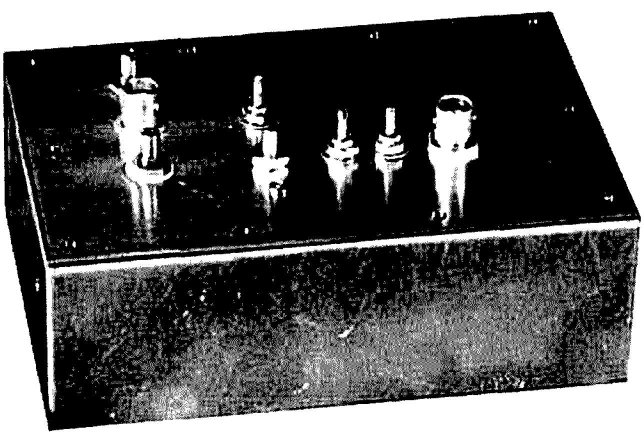
fig. 18. Layout for the 127-MHz to 382-MHz varactor tripler. Tripler is built on 1/16" brass. Hole sizes: A, 0.140"; B, 0.250"; C, 0.375"; D, 0.200".

summary

Now that we have discussed the mixer, linear amplifiers and local oscillator chains as well as spurious and intermodulation distortion we have reached a point where everything needs to be put into perspective. The eight power trains shown in fig. 9, when properly operated, give IMD figures greater than 20 dB. In each case, spurious outputs were checked with the aid of a spectrum analyzer and found to be adequately supressed (typically

greater than 40 dB for local-oscillator and image products). Power trains with two linear stages offer even greater spurious suppression.

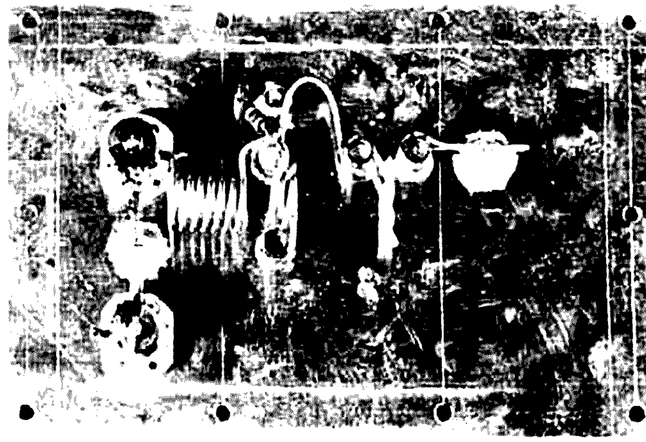
Assuming tubes of reasonable quality the eight power trains in fig. 9 have a gain margin of 2 dB; that is, the amplifiers could produce up to 2 dB less than the maximum gain in table 1 and the circuits would still function adequately. The gain margin drops slightly more than 1 dB for excitation below 30 MHz. This is due to the insertion loss of the interdigital band-pass filter. Unlike the case of class-C amplifiers, if you fall short in gain by more than the margin, the system would still deliver a valid output but at a reduced level. This is because there is no drive threshold in linear amplifiers at which output suddenly falls off.



Varactor tripler provides output at 382 MHz with 127-MHz drive.

A power train using a 3CX1000A7 triode in grounded-grid linear service is described by W6SAI. This amplifier should easily be able to run one to two kilowatts PEP input. Although this particular configuration has not been verified by actual tests, calculations indicate that the 4CX250B should be capable of driving the high-power amplifier with ease.

Needless to say, there are more expensive tubes which will serve as excellent linear amplifiers at 432. There are also tubes which will operate satisfactorily at reduced input at this frequency. However, keeping in mind the philosophy of this article, I have endeavored to select



Construction of the varactor tripler.

the most economical tubes which are not on the verge of losing steam at 432 MHz.

I would like to thank the Microwave Instruments Division of AIL for permitting the use of their newly developed microwave spectrum analyzer as well as other rf test equipment. I would also like to thank R. Kandle, K2RIW, for his comments and suggestions concerning the text. Thanks are also extended to W. Doesschate of Amperex Corporation and Bill Orr of Eimac Division of Varian for their valuable technical assistance throughout the course of this project.

references

1. James Brannin, K6JC, "SSB Converter for 432 MHz," *ham radio,* January, 1970, page 48.
2. Eimac, "An Eimac Family of 4CX600 Tetrodes," Application Bulletin no. 14, Eimac Division of Varian, San Carlos, California.
3. Reed E. Fischer, W2CQH, "Interdigital Band-pass Filters for Amateur VHF/UHF Applications," *QST,* March, 1968, page 32.
4. William Orr, W6SAI, "Two-kW 432-MHz Amplifier Using Eimac 3CX1000A7," Eimac Amateur Service Newsletter, AS25.
5. Frank Jones, W6AJF, "VHF For the Radio Amateur," Cowan Publishing Corporation, New York, 1961.
6. Harry Ferguson, K7UNL, "Linear VHF Tank Circuits," *ham radio,* November, 1970, page 56.

ham radio

solid-state carrier-operated relay and call monitor

Here are two simple but effective circuits to enhance fm operation

Murray Ronald, VE4RE, Box 974, Brandon, Manitoba, Canada

After the initial fascination of operating in the fm mode has worn off, the more technically adventurous want to add gadgets and refinements to their equipment. Preamps are popular; others go the route of tone calling, etc. Still others try the carrier-operated relay (COR), using it to trigger a tape recorder or activate a monitor for their private channel.

At this station I wanted to try some solid-state COR circuits and began experimenting with mockups using bipolar transistors (**fig. 1**). One of the problems encountered in using conventional transistors with a tube-type receiver was the undesirable loading effect introduced when the COR was attached to an i-f grid or to the squelch dc amplifier. Operation (especially in the squelch circuit) was upset considerably. By prefacing the relay driver with an fet (which has a very high input impedance) it is possible to attach the unit to the receiver without disturbing results.

carrier-operated relay

The COR shown in **fig. 2** has been tried in a number of Motorola Sensicon A and G model receivers and seems to work best when attached to the grid of the last i-f amplifier (455 kHz). Referring again to **fig. 2,** a negative voltage at point A of approximately 3.5 volts or greater will stop the fet from conducting and allow the voltage at the base of the transistor to rise. When this occurs, the transistor will conduct and pull in the relay. Typical swing in the Sensicon G receiver at the grid of the last i-f amplifier is from -1 volt with no signal to -4 volts with a 1 μV signal.

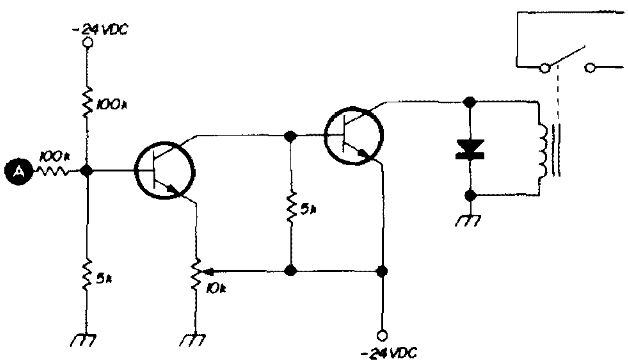


fig. 1. Carrier-operated relay using bipolar transistors. This circuit tends to load down the i-f amplifiers and squelch in tube-type receivers.

call monitor

Illustrated in **fig. 3** is a call monitor that will latch on and give an indication that a carrier has appeared on a channel. Experiments have indicated that the monitor is best attached at the grid of the dc amplifier in the squelch circuit. This is necessary because the monitor should be biased positively in the NO CALL mode to prevent noise from inadvertently triggering the 3N84. Typical voltage change in the Sensicon G at the grid of the dc amplifier is from about +3 volts with no signal to -3 volts with a 0.5 μV signal. Operation of the fet stage in the monitor is similar to that of the COR circuit. Once the 3N84 is triggered, however, it will continue to conduct and operate the Sonalert until the anode circuit is broken. A suitable low-current bulb could be used in place of the Sonalert as an indicator is desired.

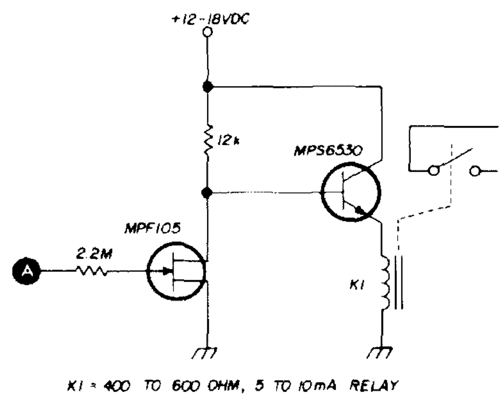


fig. 2. COR circuit using high-impedance-input fet ahead of the relay driver. Circuit works best when inserted at the grid of the last i-f in Motorola Sensicon A and G receivers.

Although the ideas in these circuits are certainly not new or original, it is hoped they will prompt some experimentation and building along these lines. Thanks go to Reg (VE4RW) for assisting in the wiring and testing of the circuits.

ham radio

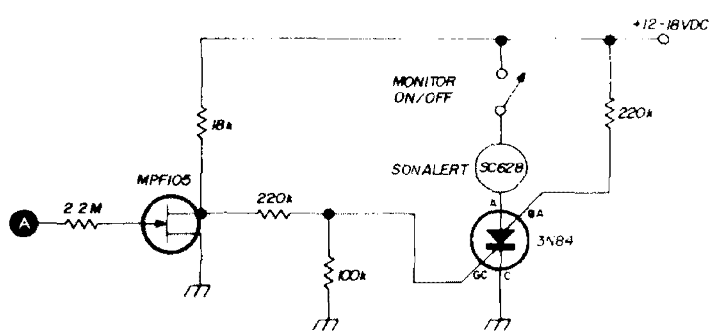
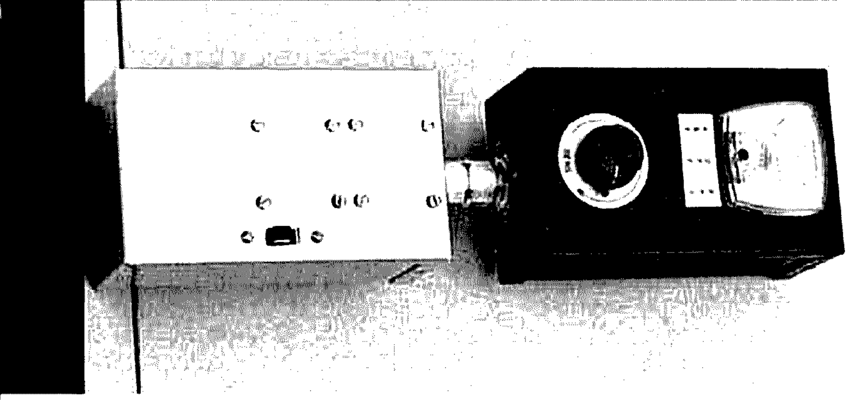


fig. 3. Call monitor that indicates when a carrier is on frequency. Experiments indicate that the monitor works best when attached at the grid of the dc amplifier in the squelch circuit.



one-man antenna matcher

A sensitive swr bridge and milliwatt signal source are featured in this compact instrument

A. G. Shafer, W4SD, 683 S. W. 7th Street, Boca Raton, Florida 33432

The adjustment of an antenna matching section (T-match, gamma, etc.) is usually a two- or three-man job – one at the antenna, one at the transmitter, and possibly a third to relay information. Because the matching adjustment involves the transmitter, some means must be used to reduce its output while carefully preventing overload of the final-amplifier tubes, particularly in many of today's rigs using TV sweep tubes. The radiation of considerable rf energy, necessary to give meaningful indications on swr meters such as the Monimatch and the wattmeter types, causes interference on already crowded bands.

The instrument described here eliminates all these problems. It's completely self-contained, weighs about a pound, and radiates only 0.1 watt maximum. One man at the antenna-matching section does the whole job; no assistants are needed, and the station transmitter is not used.

description

The instrument consists of a resistance bridge and transistor amplifier (fig. 1) and an rf signal source (fig. 2). The signal source, constructed of readily available modular units,* can be put together in minutes.

Unlike the Monimatch, this instrument isn't frequency sensitive. One hundred mW will drive the meter to full-scale deflection on 10 through 80 meters.

The bridge uses 1/2-watt composition resistors. Resistor Rs, which determines impedance, must be close to the desired value for your equipment (e. g., 52 ohms for most transmitters and transmission lines). Bridge-arm resistors R1 and R2 must be closely matched, although their exact value isn't critical.

R3 and R4 should be close in value if comparable input and output readings are to be obtained. Likewise, diodes CR1 and CR2 (1N34A's) should be closely matched.

Capacitors are disc ceramics. The 2N107 transistor has medium gain and works well with a 1.5-volt dry cell. A lower-gain device may require 3 volts to give full-scale meter deflection. Polarities shown are for a pnp transistor; an npn can be used by reversing supply polarity.

Switch S1 is a 2-pole, 3-position switch. A rotary type or a slide switch with center position off can be used.

bridge construction

The bridge, meter, and transistor amplifier are contained in a 5-1/4 x 3 x 2-1/8 inch aluminum minibox. The shielded compartment may be made from heavy aluminum or flashing copper. The shield is in the form of a

*International Crystal Mfg. Co., 10 North Lee, Oklahoma City, Oklahoma 73102. The OX oscillator is $2.95; the PAX-1 amplifier is $3.75. Both in kit form.

Z-bracket, which is attached to the end wall of the minibox. The shield shown in the photo was made from two pieces of 1-inch aluminum angle stock fastened together to form a Z.

Shielding and parts assembly of the bridge are important. The only elements within the shield compartment are resistors Rs and R2, which are mounted at right angles to each other. A piece of RG-58/U cable is connected between the SO-39 chassis connector at the opposite end of the box and an insulated stud within the shielded compartment. The center conductor is connected to the stud and the shield braid to a ground lug next to the stud.

Connections to the remaining bridge elements are made through clearance holes in the rear wall of the shield. Short leads are required up to and including the point at which R3 and R4 are connected. The usual soldering precautions are suggested, and the use of heat sinks (long nose pliers, etc.) is recommended. The layout of the other parts is not critical and follows a logical sequence. If the box size and layout shown are followed, be sure the meter doesn't extend inside the box by more than 7/8 inch exclusive of terminals. Note that the SO239 output connector and the insulated stud are centered close (11/16 inch) to the open side of the minibox.

After assembly and wiring, and with the battery in place, it will be noted that a small reading (a few microamps) will be indicated on the meter with the switch in either the FWD or REF position. This is the "no-signal" current of the transistor and is the "zero" indication of a perfect match when the bridge is in use.

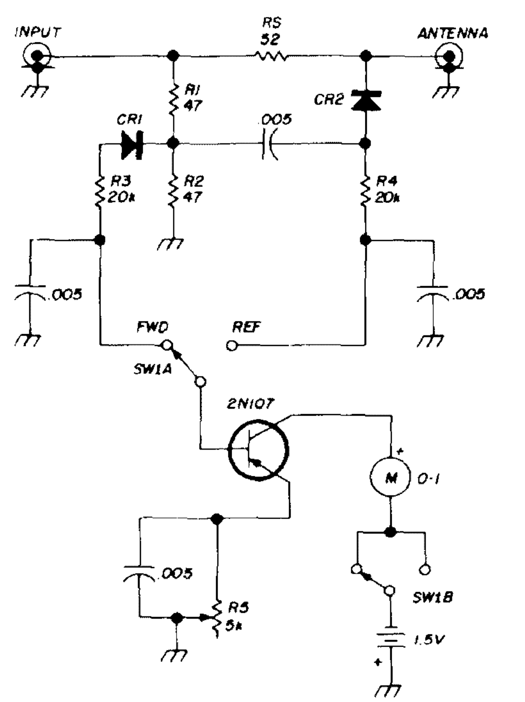


fig. 1. Schematic of the bridge circuit and transistor amplifier. Matched pairs are recommended for R1, R2; R3, R4; and CR1, CR2 (see text).

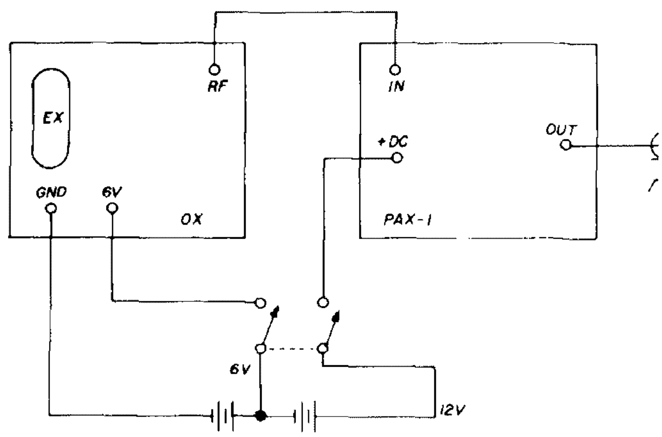


fig. 2. Signal source and amplifier. Units are available in kit form; easily assembled.

signal source

The OX Oscillator uses an EX Crystal of the frequency at which the antenna matching is to be accomplished. The OX – PAX-1 combination can be made to function over two adjacent ham bands by changing crystals and repeaking the coils. The OX – PAX-1 will put out up to 200 mW, which is more than enough to drive the bridge. The units are assembled in the same size minibox as the bridge. A much smaller box would accommodate the signal-source units, which are only 1-1/2 inch square, but it was desired to mount the battery, consisting of eight penlight cells inside the box, requiring the extra space.

Note that a *double*-pole switch is used to turn off both the 6- and 12-volt lines. Disabling only the negative line will result in a battery drain in the OFF position (between the 6- and 12-volt taps).

The rf generator may be connected to

the bridge by a double male adapter or by a length of RG-58/U cable if the rf unit is placed in your pocket.

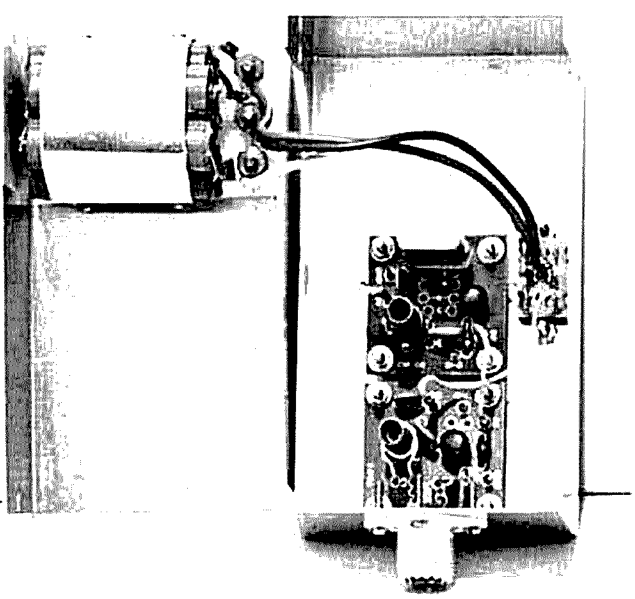
The connection between the bridge and the antenna matching section must be as short as possible. The matching section can terminate in a female connector such as the SO-239, or a couple of inches of RG-8U and a PL-259 could be used. If the SO-239 is used, the bridge should be connected to it by a double male adapter.

checkout

Before using the equipment for matching or swr measurements, the bridge should be tested. Solder a 52-ohm resistor into a PL-259 connector as a dummy load, and insert it in the output connector of the bridge.

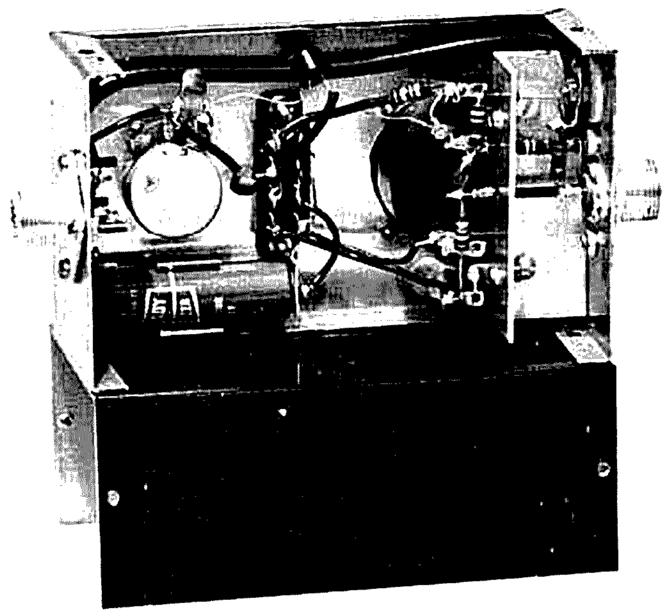
With the rf generator connected to the input of the bridge, position the switch to

Signal-generator portion showing International Crystal Co. kit modules. Power supply consists of 8 penlight cells.



FWD, and adjust the sensitivity control for a 1-mA reading on the meter. Positioning the switch to REF should reduce the meter reading to zero. If the meter doesn't drop to zero, bridge arms R1 and R2 are not equal, Rs is not 52 ohms, or coupling exists between the bridge elements.

If this test is satisfactory, remove the dummy load, and with the output open-circuited, adjust the meter reading for 1 mA in the FWD position. With the switch in the REF position, a 1-mA reading should be obtained. Repeat this last test



Bridge and meter portion of the antenna matcher. Shielding and parts placement are important. Power is supplied by a 1.5-volt dry cell.

with the output shorted by a *very* short wire, resetting the meter to 1 mA in the FWD position. It should read 1 mA in the REF position.

Slight resistor and diode variations may make such correlations not quite as exact as indicated, but such differences should not exceed a few microamps.

calibration

If accurate swr measurements are to be made, the meter should be calibrated by using resistors of two, three, and four times the value of bridge resistor Rs. These resistors should be soldered into a PL-259 connector for such calibrations.

acknowledgement

I want to express my sincere thanks to my SWL friend and potential ham Herwart Werker for the photos.

ham radio

audio agc

principles and practice

Audio agc has many uses in amateur equipment — this article describes how it works and presents a practical circuit

Courtney Hall, WA5SNZ, 7716 La Verdura Drive, Dallas, Texas 75240

This article should provide you with sufficient information to build a simple audio agc circuit, understand its operation, and integrate it into your particular application.

Agc stands for automatic gain control; audio agc is a means of equalizing weak and strong signals and prevents overload and distortion in the agc amplifier and following stages when strong signals appear at the input. **Fig. 1** shows a widely used agc circuit in block form. Signals pass through the control element and are amplified to a level sufficient to be detected; the detector output is a dc voltage which increases when the input signal gets larger and decreases when the input signal gets smaller. This dc control voltage is fed back to the control element which attenuates the signal in proportion to dc control-voltage amplitude. The overall effect is that the output signal amplitude remains relatively constant as the input signal amplitude varies over a wide range, thus providing high gain for weak signals and low gain for strong signals. Manual gain control should always *follow the agc circuit* to adjust its relatively constant output to a level suitable for the following circuits.

applications

One of the most useful places for audio agc is between the microphone and *transmitter*. The agc circuit not only prevents overmodulation when speaking loudly into the microphone, it also minimizes decreases in modulation when voice level drops or the microphone is moved away. Once the manual gain control is set, no further attention should be required.

The current interest in direct-conversion receivers has brought another application for audio agc to mind. Several articles have been published describing these simple receivers, but few, if any, include any form of agc. Since most, if not all, of the gain of these receivers is provided in the audio amplifier, this would be the logical place for an agc

circuit. Audio agc would prevent distortion and ear fatigue caused by strong signals.

Other uses include telephone amplifiers and tape recorders. Audio agc is also a deterrent to audio *howl* or *feedback* common in public-address systems; the howl appears as a signal to the agc circuit which then reduces gain to control it.

agc vs clippers

Speech clippers are not uncommon in transmitters and usually include two diodes which clip the positive and negative excursions of the audio signal when its amplitude reaches a certain level. Such clipping circuits can prevent over-modulation, but they have two shortcomings compared to audio agc. When a speech clipper is limiting the audio signal, it is also distorting it and producing harmonic frequency components not present in the original speech signal; this distortion decreases intelligibility. Filters following the clipper do not remove harmonics of the lower speech frequencies which fall within the passband of the filter. In addition to distortion, the clipper circuit provides no increase in gain if the audio signal falls to an unusually low level; this does not improve intelligibility either.

Audio agc circuits do not have to introduce distortion to perform their function. They can be designed to hold output constant within a few dB while the input varies over a range of 60 dB or more.

You may think that clipping circuits have the advantage of simplicity but this argument is rather thin. Both techniques require an amplifier with enough gain to bring the signal up to the level required to forward bias a diode. The agc circuit requires no low-pass filter for the signal but needs a detector and control element. The advantages of a clean signal with *almost constant amplitude should* be weighed against the slight increase in complexity of an audio agc circuit.

agc characteristics

There are practical limits to the performance of any circuit, and conflicting requirements often dictate compromises in design goals. The following paragraphs describe the important characteristics and limitations of audio agc.

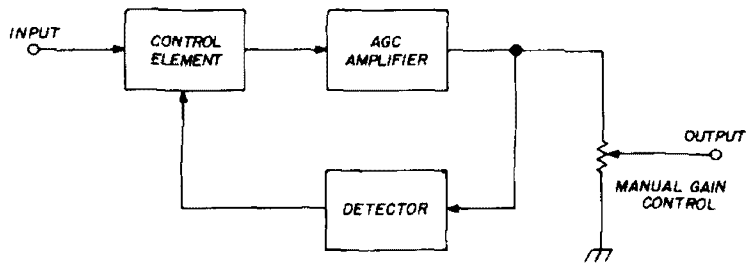


fig. 1. Block diagram of a basic agc circuit.

At extremely low input signal levels, the agc amplifier does not have sufficient gain to cause the detector diode to conduct; therefore the dc control voltage is zero, and the control element does not attenuate the signal. As the input signal is increased the agc amplifier output voltage increases linearly until the detector diode begins to conduct and produce a dc control voltage; this point is called the agc threshold because it is the point at which agc action begins.

Input signals below the agc threshold are amplified linearly, and input signals above the threshold are amplified or attenuated as needed to hold the output voltage constant. The agc amplifier gain can be increased until the threshold is so low that agc action occurs with circuit noise. This assures agc action on the lowest usable signal, but the signal-to-noise ratio of larger signals will be seriously degraded.

If amplifier gain is increased without discretion the dynamic range of the

control element may be exceeded with only moderately strong signals, causing overload and distortion. Amplifier gain must be selected so that the threshold is at the optimum point with respect to expected input signal levels and acceptable signal-to-noise ratio.

Attack time is the time required to reduce the gain when a strong signal suddenly appears at the input. It is important that attack time be relatively fast so that gain can be reduced before distortion occurs. This parameter is highly dependent on the charging time of the filter capacitor in the detector circuit.

Release time is the time required to increase gain when a strong signal is suddenly removed from the input. This time is relatively long, on the order of one second, so that gain does not fluctuate between words and syllables. Release time is controlled primarily by the discharge time of the detector's filter capacitor.

Distortion is an important parameter in any audio system. A well designed audio agc circuit should not show any significant distortion of output waveform when viewed with an oscilloscope.

practical circuit

There are many configurations and variations used to accomplish audio agc. Discussion of all these techniques is beyond the scope of this article, so attention will be focused on one type of circuit which I consider to be easily reproducible.

Bipolar and field-effect transistors have wide variations of parameters in devices of the same type. However, the forward-bias characteristics of silicon diodes are relatively uniform. Therefore, diodes are desirable for use as control elements if circuit reproducibility without selected devices is important.

Fig. 2 shows the forward voltage vs current for a 1N914 silicon diode. At forward bias levels less than 0.3 volt the diode is essentially off and has a very high

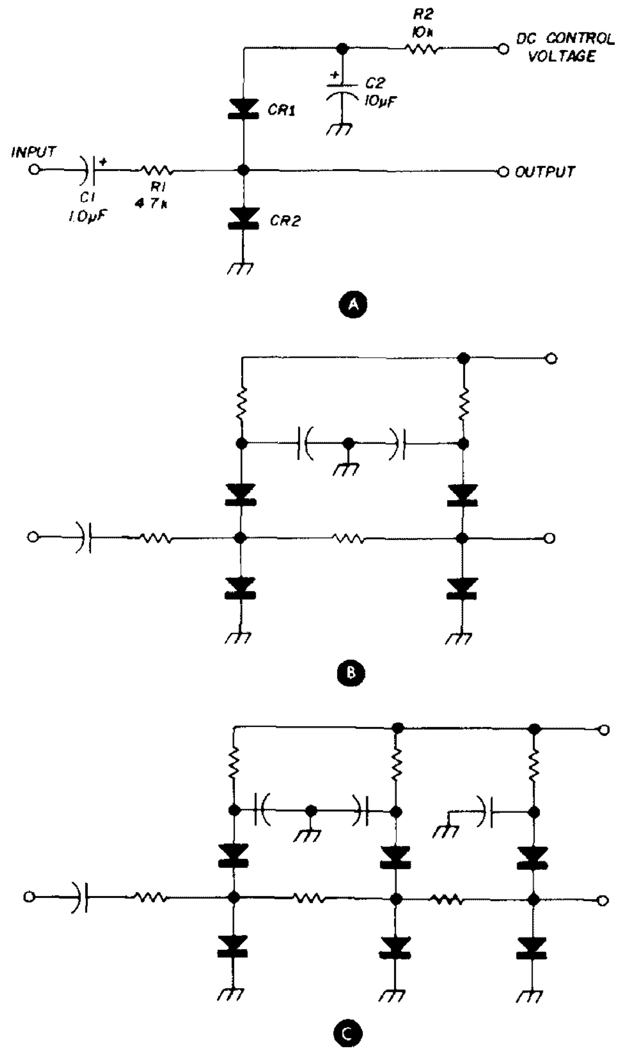


fig. 3. Voltage-controlled attenuators. A shows a one-section attenuator; two- and three-section attenuators are shown in B and C.

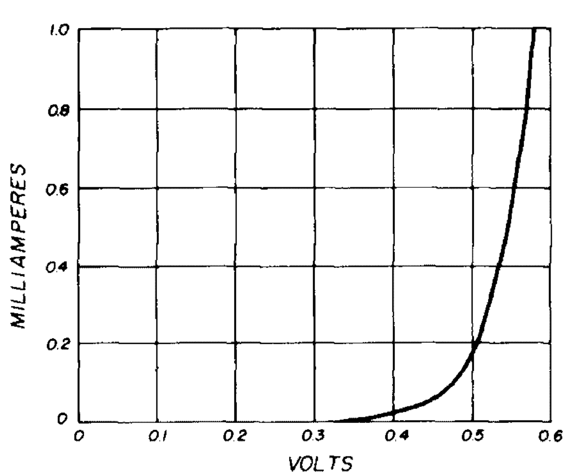


fig. 2. Forward characteristic of a 1N914.

resistance. As forward bias voltage is increased, current begins to increase more and more rapidly, and the diode *exhibits* less and less resistance. Thus, the diode

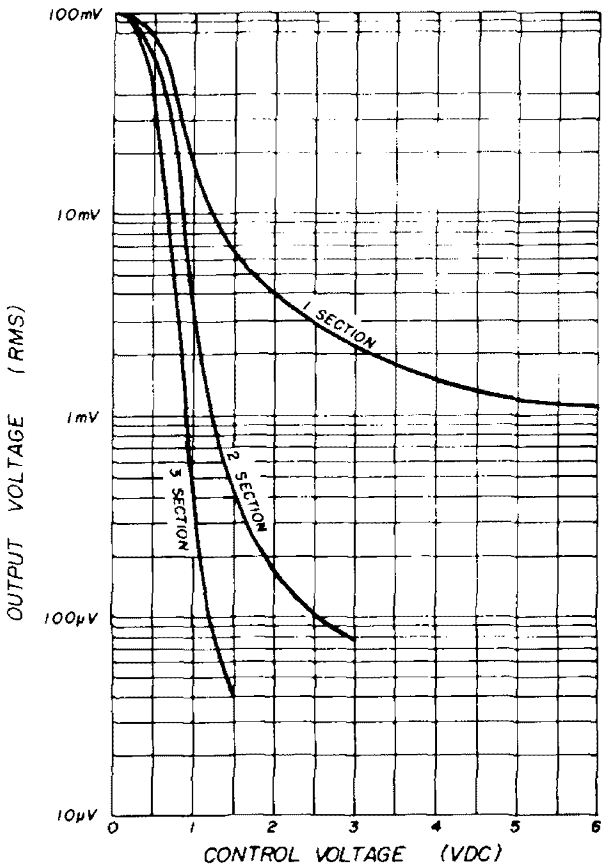


fig. 4. Response of one-, two- and three-section voltage-controlled diode attenuators.

can function as a voltage- (or current-) controlled resistor.

If the amplitude of the ac signal across this voltage-controlled resistor is kept small, resistance changes due to signal amplitude will be small, resulting in low distortion. The resistance shown by the diode to a low-level ac signal is called the dynamic resistance and is the reciprocal of the slope of the curve at any point. The slope of the diode's curve at 0.45 volt bias is approximately 1.1 milliampere-per-volt, and the dynamic resistance is about 910 ohms; at a forward bias of 0.55 volt, the slope and dynamic resistance are about 11 milliamperes-per-volt and 91 ohms, respectively.

A schematic of a single-section voltage-controlled attenuator using two 1N914 diodes is shown in **fig. 3A.** R1 acts as the series element of an L attenuator; CR1 and CR2 form the shunt element. C1 prevents control current from flowing in the input circuit; C2 bypasses the current limiting resistor, R2, out of the signal circuit. No dc control current should be allowed to flow in the output circuit.

Figs. 3B and **3C** show two- and three-section attenuators, all sections being the same. Data taken on these attenuators is plotted in fig. 4. The input signal from a 1-kHz 600-ohm generator was held constant at 100 millivolts rms, and output signal voltage was plotted vs dc control voltage (from a power supply). Output waveform was monitored on an oscilloscope, and no significant distortion was detected.

The single-section attenuator appears to approach a limit of about 40 dB attenuation, indicating a minimum shunt resistance on the order of 50 ohms. More than 60 dB of attenuation is available from either the two-or three-section circuit. The two-section attenuator was judged to offer the best compromise between performance and number of components. Accurate readings below 100 microvolts were difficult because of noise.

If the two-section attenuator is followed by an amplifier having a voltage gain of 1000 (60 dB) the overall gain of the composite circuit could vary from less than one up to 1000, depending on the amplitude of the dc control voltage. **Fig. 5** is a schematic diagram of a complete audio agc circuit. Q1, Q2 and Q3 make up the amplifier portion; CR5 is the detector. Q4 is a dc amplifier and C9 is the agc filter capacitor. The ratio of R9 to R7 determines the closed-loop voltage gain of the amplifier, which is 1000. If R7 is shorted the open-loop gain is about 56,000 (95 dB).

Amplifier bandwidth extends from 150 Hz, determined by C4, to 15 kHz, determined by C6. Other bandwidths can

be obtained by changing the values of these capacitors; the usual 300 to 3000 Hz communications bandwidth is obtained by using .01 μF for C4 and .05 μF for C6. R5, R10 and R11 provide dc bias stabilization, and C5 prevents signal feedback via this path. Power supply drain is about 7 milliamperes. The transistor types in parenthesis are epoxy devices which should perform as well as the *hermetically-sealed types.*

μV of input voltage.

Attack time is in the neighborhood of 40 milliseconds; release time is on the order of one second. Release time can be increased by raising the value of C9 and attack time can be decreased by reducing R12, but there will be some interaction in these adjustments. Some experimentation

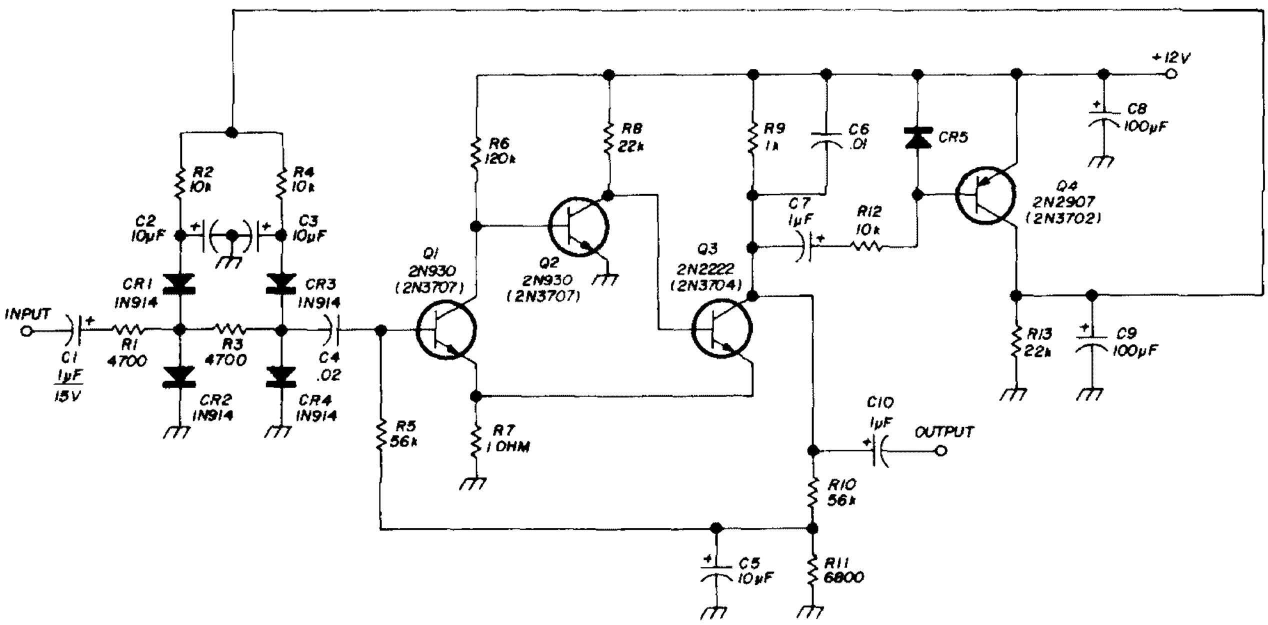


fig. 5. Schematic diagram of a practical audio agc circuit. Complete performance characteristics are shown in table 1. Input/output characteristic is plotted in fig. 6.

The input impedance of the amplifier, looking from C4, is about 56,000 ohms; this resistance and the two series resistors in the voltage-controlled attenuator form a voltage divider which reduces the maximum voltage gain of the circuit to about 860.

Test data on the circuit of **fig. 5** is listed in **table 1.** The ratio of maximum to minimum gain is 1430, and the maximum amount of harmonic distortion was measured at less than 2.5%. Data is omitted where the signal was too small to be measured with reasonable accuracy. Input voltage vs output voltage is plotted in **fig. 6.** This graph shows that agc threshold occurs at approximately 300

should disclose the optimum values for these components. I have observed that when attack time is decreased below a certain point the circuit oscillates at about 1 Hz. This behavior has not been investigated to my satisfaction, but it is corrected by increasing R12, or by raising the amplifier's lower cutoff frequency (decreasing C4).

using the agc circuit

Successful incorporation of this circuit into an audio system depends heavily on proper interfacing at the input and output. If the agc circuit is inserted between the microphone and transmitter an attenuator should be placed between the

agc output and the transmitter's mike input. The maximum agc output, nominally 0.5 volt, should be reduced to the order of one millivolt to provide normal input level to the transmitter.

Circuits connected to the agc output should not appreciably load the 1000-ohm output impedance, or the gain will be lowered. Assuming typical microphone output varies from 30 μV to 10 mV as sound level changes then a large portion of the lower level sounds would not be within agc range. If more complete sound leveling is desired, a 20-dB preamplifier could be connected between the microphone and the agc circuit's input. This would have the effect of decreasing the input voltages in fig. 6 by a factor of 10, and the new threshold would occur at a microphone output of 30 μV. Assuming the preamp had the same equivalent input noise as the agc circuit the noise output from the agc circuit would be about 10 times higher or 22 mV. This would still provide a minimum signal-to-noise ratio of more than 20 dB in the agc range.

If the audio agc circuit is used with a simple receiver the receiver's audio should be amplified or attenuated to a level consistent with good agc action and acceptable signal-to-noise ratio before feeding into the agc circuit. The receiver volume control should be connected between the agc output and the receiver's output amplifier.

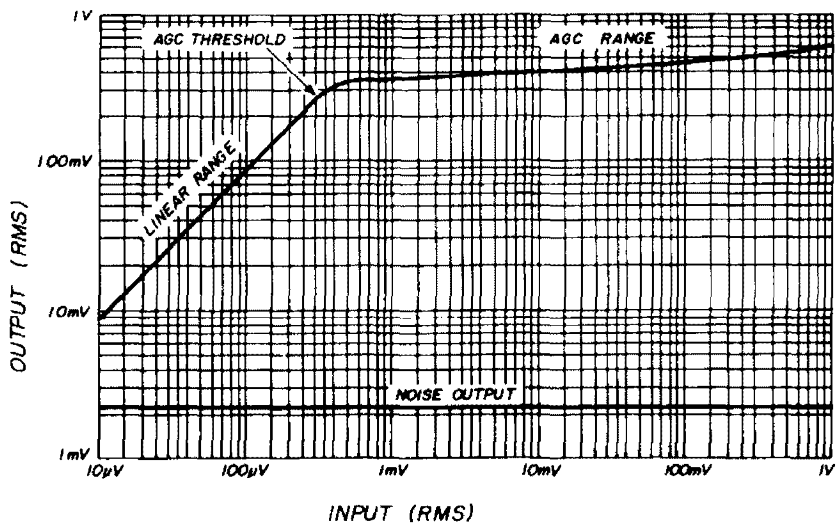


fig. 6. Input vs output of agc circuit shown in fig. 5.

A useful item to include in the agc circuit is a meter to monitor dc control voltage; this would provide a visual indication that signals are within the agc range. Such a meter circuit should not load the detector dc amplifier. It has been found satisfactory to use a 50-μA meter in series with a 100,000 ohm resistor as a 5-volt full-scale meter connected between the collector of Q4 and ground.

table 1. Performance of the agc circuit of fig. 5.

input voltage (rms)	output voltage (rms)	voltage gain	harmonic distortion (%)	dc control voltage (volts)
10 μV	8.6 mV	860	—	—
32 μV	27.4 mV	855	—	—
100 μV	85 mV	850	—	—
320 μV	270 mV	843	—	0.04
1 mV	360 mV	360	0.9	0.65
3.2 mV	380 mV	119	0.9	0.81
10 mV	400 mV	40	0.9	0.98
32 mV	430 mV	13.4	1.2	1.19
100 mV	460 mV	4.6	1.2	1.51
320 mV	520 mV	1.6	1.3	2.03
1 V	590 mV	0.6	2.4	2.95

conclusion

This article is not intended to be an exhaustive study of audio agc theory and technique, but it is hoped that it will assist those readers who wish to experiment in this area.

ham radio

fixing a sticky AR-22 rotator

Most sticking AR-22 rotators are caused by the same problem — easily fixed with three rivets

Ted Woolner, WA1ABP, 30 Cedar Road, Shrewsbury, Massachusetts 01545

Many amateurs use the CDE AR-22 rotator with great success, especially where there is little cold weather or icing. However, in areas subject to ice storms, operators may experience trouble with a sluggish rotator, or one that only goes part of the way around. When the AR-22 is strained by heavy loading, the rather husky motor tends to bend over the teeth in the drive gears.

The three drive gears, part number TRA-39, consist of three thin iron gears sandwiched together on one spindle. They drive the TRA-18 ring gear which goes around the perimeter of the rotator. If one of the gears in the sandwich happens to be a little larger, it will take all the strain and can wear to the point where the other two will wear unevenly. When all three gears are sufficiently worn, they start to bind. This usually starts at one particular point in rotation, but eventually spreads to the entire 360° and may cause the aluminum ring gear to snap in two.

The cure for this malady is rather simple: rivet the three stamped gears together. This way one gear won't take all the loading, and the three gears will wear evenly. Remove the TRA-18 ring gear and TRA-39 drive gears (use your instruction book for guidance). If your rotator has been binding, order a new TRA-39 drive gear from CDE.* Even a slight bend in these gears will lead to eventual trouble.

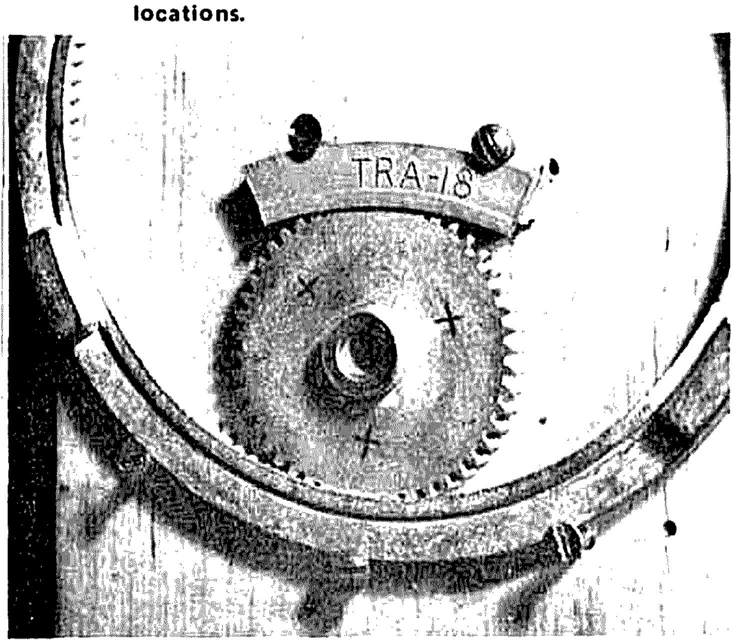
*Send all parts orders to Cornell-Dubilier Electronics, Rotor Parts Department, Desplaines, Illinois 60018, *not* to their factory in Fuquay Springs, North Carolina.

Also inspect the TRA-18 ring gear for damage. Wash it off in solvent and look carefully for hairline cracks. This is also an opportunity to look over the rest of the parts in the rotator; if any are broken or badly worn, replace them.

Since each of the gears in the TRA-39 assembly was stamped out, each has a slight burr on one side. Do not try to file them flat. When mounting them on the spindle be sure to place them with the burrs toward the bottom; otherwise the gear stack will be too thick.

To drill the rivet holes in the TRA-39 gear, mount the ring gear on a piece of wood with screws as shown in the photo. Mesh the TRA-39 with the ring gear and hold it in place with a section of damaged ring gear. The beveled heads of the wood screws will force the gears together and hold them firmly in place. Drill three equally-spaced holes in the TRA-39 drive gear. If you have a drill press, use it; it will insure that the holes are perpendicular to the face of the gears. Use soft-iron rivets to hold the gears together; rivets 3/8-inch long are just about right. The rivets should be snug in the drill holes for maximum strength.

To drill the rivet holes, the TRA-39 drive gear is held down with a TRA-18 ring gear and sawed-off section from a damaged ring gear. The small crosses mark the rivet-holes locations.



Remember when reassembling that the ring gear should be placed in position last. Make sure that the line stamped on the cam gear is parallel with the edge of the motor-mounting plate as shown in fig. **1.** In this *position* the pulsing-switch points should open; rotate the gears until this happens. The stop arm should be pushed to the left (counterclockwise) as far as it will go. In this position the stop lug on the ring gear should be up against the stop arm.

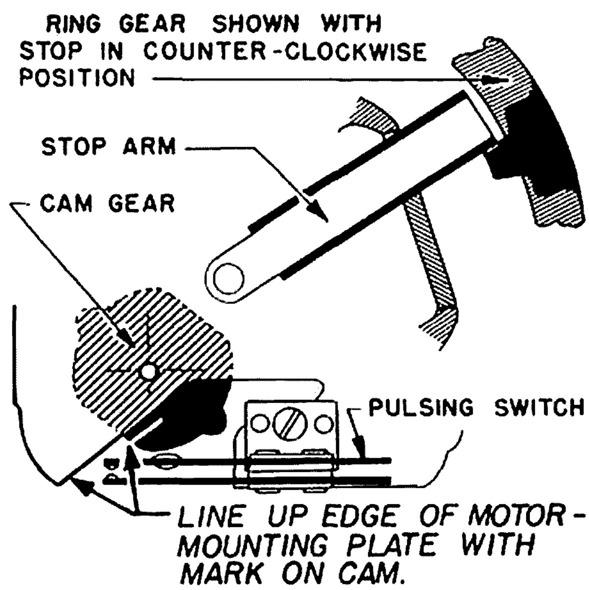


fig. 1. When putting your AR-22 back together, the line on the cam should line up with the edge of the motor-mounting plate as shown here. The ring gear should be against the stop arm in the counterclockwise position.

Be sure that all the ball bearings are in place in the retainer spring clips. Thoroughly grease the gears with a good silicone grease such as Dow-Corning 44 or Lubriplate, a white lubricant available in many hardware stores; an 8-ounce tube is sufficient. Also, when replacing the top cover of the rotator, make sure the lugs in the ring gear rest in the recessed sockets provided inside. When the rotator is assembled in this manner it is in the *North* position, against the stops from a counter-clockwise direction through *East*. The control box should be oriented to coincide with this setting.

ham radio

the electronic hand keyer

An ordinary hand key or bug can be used with this circuit to form perfectly shaped code characters

H. Paul Clampit, K5TCK, 1125 Ridgeview, Mesquite, Texas 75149

Today's electronic keyers are controlled by some type of paddle with two sets of contacts – one for dots, the other for dashes. The "electronic hand keyer" is controlled by only one set of contacts. A standard hand key or a semiautomatic key (bug) will work perfectly. Dots or dashes can be made, and the spacing and ratio will be perfect. Suddenly your fist will sound like a million dollars!

operation

The electronic hand keyer will generate a dot or dashes depending upon how long the key is held closed. If the key is released before the correct dot interval is over, a perfect dot results. If the key is held down longer, a dash will be made. If the key is held down still longer, additional dashes or an additional dot can be made.

For example, to make a 9 requires only one press of the key. The key is pressed until four dashes are made and then released during the first one-third of the fifth dash, thus making it a dot. For characters similar to 9 the electronic hand keyer is superior to the fully automatic twin-lever keyer. Unfortunately, a study of the character 4 quickly puts the electronic hand keyer back into its place.

The key must be released quickly to make a dot, or else a dash will result; so to make a 4 requires the character to be sent just as it would be with a hand key. The electronic hand keyer essentially takes the 4, as sent by the hand key, and corrects the spacing and dot-to-dash ratio. A character sent by a bug is corrected in the same manner. All characters can be sent normally by a hand key or bug to be corrected, or reshaped, by the electronic hand keyer.

basic keyer

The circuit is essentially that of a simple electronic keyer with one connection changed. Only the basic function of the electronic hand keyer is discussed in this article. Detailed circuits with component values are not given. The electronic hand keyer is presented in this manner because of subparagraph 807 of Murphy's Law.* You are encouraged to study and understand the function of the circuit, then design and construct your own from available components.

A simple keyer is shown in **fig. 1.** The clock can be a free-running multivibrator, a unijunction relaxation oscillator, or almost any adjustable source of low-frequency periodic signal. FF1, FF2 can be any triggered flip-flop connected as a divider so that it will change state each time the negative (trailing) edge of the trigger signal is received. The Clear terminals (C1, C2) hold the Q terminal low as long as the Clear signal is high. A relay output is shown, but a keying transistor could easily be used instead.

The clock runs continuously, and both flip-flops are normally biased off (clear).

*Subparagraph 807 of Murphy's Law states that "The reader's junk box will *never* contain the components required by the magazine article."

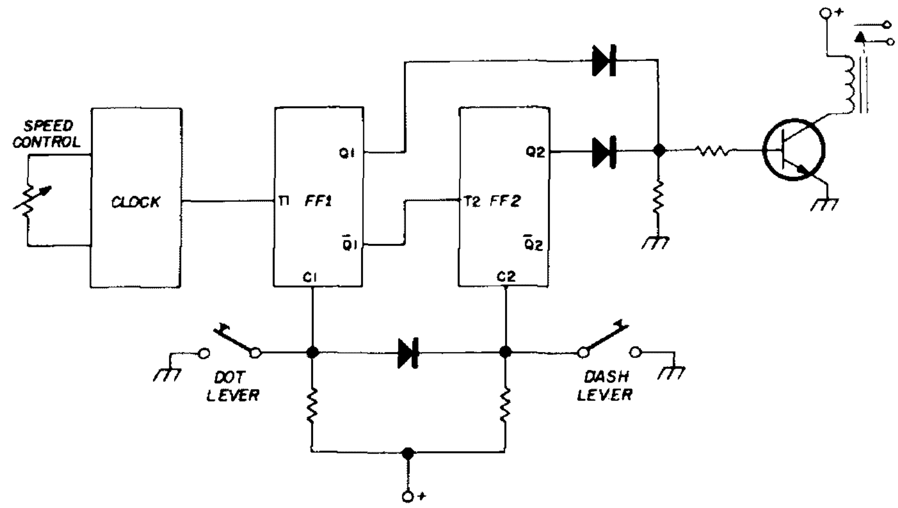


fig. 1. Basic keyer circuit. The clock is a free-running oscillator. After the dot or dash lever is held closed, a negative edge from the clock triggers the flip-flops thus forming a character.

A character is started only after the dot or dash lever is held closed until a negative edge comes from the clock. **Fig. 2** shows the signals that occur when the dash lever is closed and held closed beginning during some time interval, A. A diode pulls C1 low when C2 is forced low. The diodes connected to Q1 and Q2 serve as an OR gate to pull in the relay when either Q1 or Q2 is high, thus making a dash.

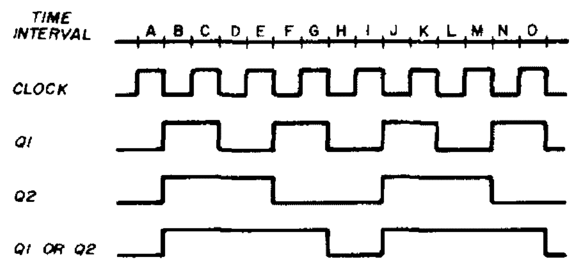


fig. 2. Timing diagram of dashes made with the circuit of fig. 1.

self-completion

The circuit of **fig. 1** does not self-complete because the flip-flops are forced clear immediately when a lever is released, allowing the Clear terminals to go high. The character in progress will then be chopped off immediately if the lever is released.

Fig. 3 shows this same keyer with two diodes added to make the keyer self

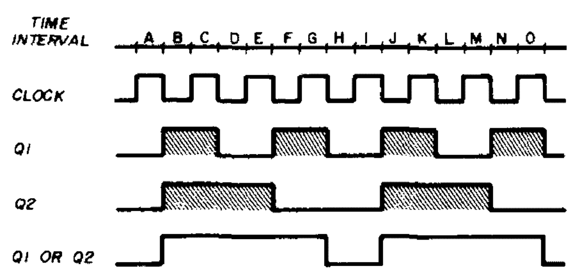


fig. 4. Timing diagram for circuit of fig. 3. Shaded areas show when flip-flop Clear terminal is held low by its associated Q̄ terminal.

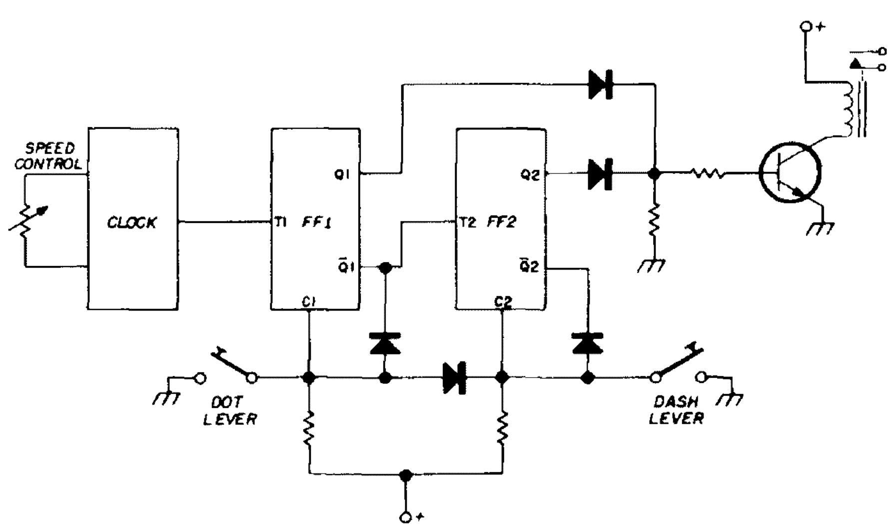


fig. 3. Keyer circuit of fig. 1 with two diodes added to make the keyer self-completing.

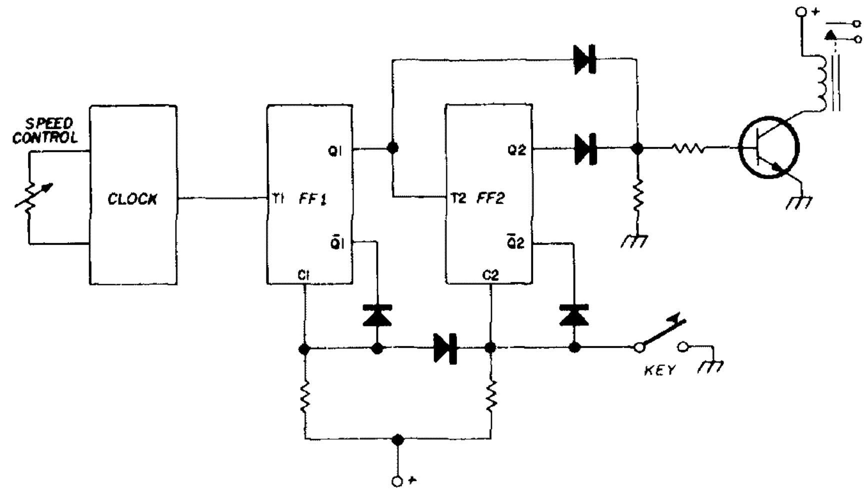


fig. 5. Circuit of the electronic hand keyer.

completing. The Clear terminal is now pulled low by the diode if the $\overline{Q}$ terminal is low. The shaded areas of **fig. 4** show the times when the Clear terminal is held low by its associated $\overline{Q}$ terminal. When a dash is started the dash lever can be released, and $\overline{Q}2$ holds the dash Clear terminal low for two-thirds of the dash; then $\overline{Q}1$ holds the dot Clear terminal low for the remainder of the dash.

Notice that Q1 must go high before Q2 goes low or the self completion will be lost, and the dash will be terminated when it is only two-thirds complete. Fortunately, since FF2 is triggered by FF1, this condition is always met.

the electronic hand keyer

If FF2 is triggered from Q1 rather than $\overline{Q}1$, as shown if **fig. 5**, the timing diagram of **fig. 6** results. The shaded areas again indicate intervals where the Clear terminal is held low by the associated $\overline{Q}$ terminal of the flip-flop. Notice that during intervals C, D and K, L Q1 goes low before Q2 goes high. If a dash is started and the lever released, the self-completion is lost when Q1 first goes low, and the dash is terminated when it is only one-third complete. If the dash lever is held for over one-third of the dash interval, $\overline{Q}2$ takes hold, and the full dash will be made.

summary

Fig. 7 shows the shaping capability of the electronic hand keyer. Poor code is reshaped perfectly if it is sent at or near the same speed as that to which the circuit is adjusted. To send a dot or a dash, it is necessary only to hold the key down until the character starts, then release it at the proper time.

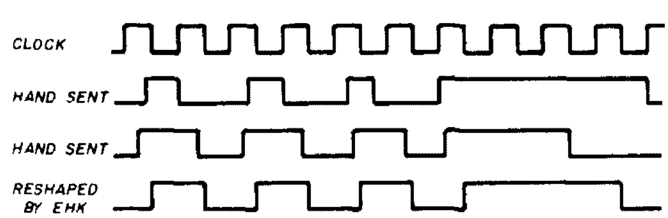


fig. 7. The shaping capability on the letter "V" by the electronic hand keyer.

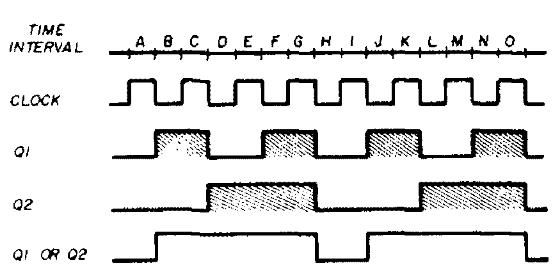
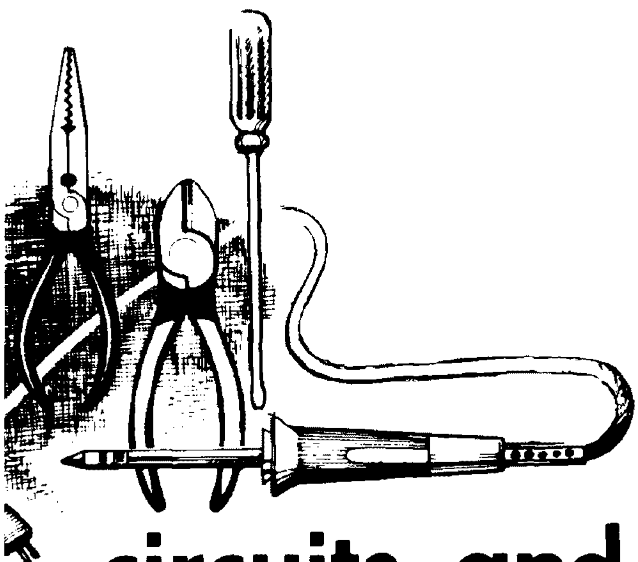


fig. 6. Timing diagram showing self-completion of dots and dashes of the electronic hand keyer.

The electronic hand keyer will also make an interesting and educational project for the beginning digital and solid-state experimenter. The foward-looking amateur might want to add a switch at the T2 terminal to change the electronic hand keyer back to a regular keyer once he has progressed that far.

ham radio



circuits and techniques

ed noll, W3FQJ

integrated circuits

The foreboding opinion that integrated circuits will stop amateur experimentation and stymie ingenuity is unfounded. In fact, so much reliability and versatility have been built into these devices that there appear to be an infinite number of external circuits and systems yet to be tried. Each amateur can look forward to a lifetime of fun and experimentation with solid-state devices and systems; the integrated circuit is just an extension of the solid-state science of packing active devices into ever smaller spaces. Diodes, transistors and resistors are the primary components used in integrated circuits although a limited number may include an occasional capacitor or coil.

Since capacitors take up considerable space it is customary to use circuits that do not require capacitance. Also, it is difficult to design a precise value resistor into an integrated circuit. On the other hand, there is no great problem in including two or more resistors of exactly the same value even though a *certain* absolute value is difficult to attain. Hence, internal circuitry uses balanced configurations that require equal-value resistors but are not critical as to absolute value. All of this boils down to the fact that the most common integrated circuit is the balanced dc amplifier.

basic differntial amplifier

The differential amplifier is the mainstay of integrated circuits. It is basically an emitter-coupled configuration (**fig. 1**); as a dc amplifier it has fine stability and good rejection of undesired signal components. Since it is a direct-coupled amplifier no interstage coupling capacitors are needed.

Ideal differential operation requires that the two collector resistances be the same and the characteristics of the two transistors be identical. In terms of discrete component circuits this is a disadvantage because perfectly matched transistors and resistors are necessary. However, in ic production these conditions are met quite readily and at low cost. In basic operation the differential amplifier emphasizes the signal difference that exists between base inputs, developing equal-amplitude and out-of-phase collector signals.

It is stated that a *differential-mode* input signal is applied. In practice this is done by applying the desired ac signal to just one of the base inputs. Since no signal is applied to the opposite base the difference voltage between the two equals the magnitude of the signal applied to the one base.

When two equal-amplitude similar-polarity signals are applied to the base inputs the ac signals across the common emitter resistor are subtractive. When in perfect balance the differential amplifier

performs in bridge-like manner — there is no output observed from collector to collector and very reduced output from each collector and common. Such an applied signal is referred to as *a common-mode* input signal. This is usually the form of undesired signals such as hum and interference.

In the difference-mode operation a signal applied to base 1 appears at the collector output of transistor 1 and also across the common emitter resistor. The latter signal component serves as the input signal for transistor 2. As a result the output at collector 2 is opposite from that at collector 1. The differential amplifier acts as a phase splitter, developing two equal-amplitude but opposite-polarity signal components at the output.

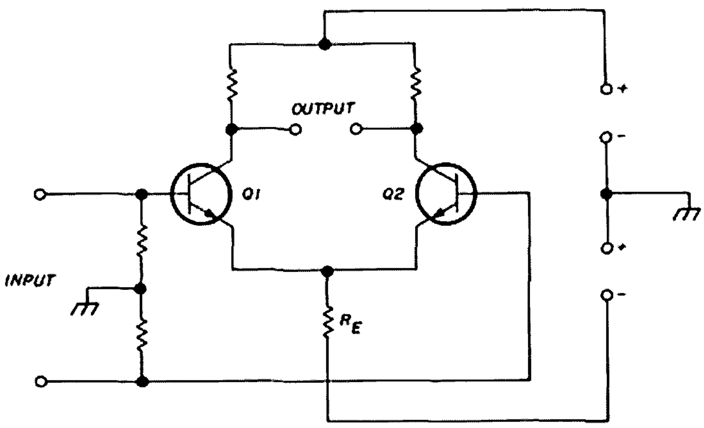


fig. 1. Basic differential amplifier.

The differential amplifier has a high order of dc stability, reducing the influence of supply voltage changes, temperature, etc. It is even practical to construct a multistage affair using the difference concept. A differential amplifier or a group of them connected in cascade arrangements are the most common circuit configurations built into integrated circuits.

In the differential amplifier, not only are the interstage coupling capacitors eliminated, the emitter bypass capacitors are eliminated as well. In making a comparison between ics and discrete circuits it should be noted that an integrated circuit has fewer passive components (resistors, capacitors and coils) and more active components (transistors and diodes) than a comparable amplifier built of discrete active and passive components.

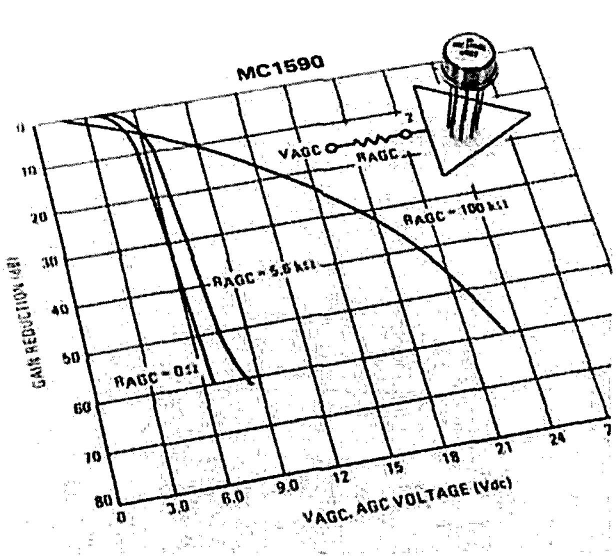
stability

In a perfectly balanced differential amplifier there is stable amplification with changes in dc operation conditions and temperatures. A change in leakage current and/or gain in one side of the differential circuit is balanced out by a like change in the second side. Such balance, and the ability to compensate for any imbalance, sets the operating limits of the differential amplifier.

Reduction of common-mode signals depends upon the degenerative effects of the common-emitter resistor. Of course, the higher the ohmic value of this resistance, the greater the rejection. Such increase is limited by supply voltage requirements and the greater difficulty of including high-value integrated resistors.

The answer to this problem is to include a constant-current emitter source composed of an additional active component, rather than a high value resistance. The fundamental arrangement is

Performance of the Motorola MC1590 integrated circuit.



shown in **fig. 2**. In this circuit the combination of the transistor and its low value emitter resistor acts as a high-resistance constant-current source. The presence of a common-mode signal on the differential transistors affects base voltages and junction resistances. However, emitter and collector currents are held constant by the constant-current emitter source. In fact, the undesired voltage change appears totally across the constant-current source, which is highly degenerative. Thus, the differential gain of the amplifier in terms of common-mode signals is greatly reduced.

The diode in the base circuit of the constant-current source provides temperature compensation. Exact compensation is obtained when the characteristics of the base-emitter junction of the constant-current transistor and the diode junction are identical. With a rise in temperature there is an increase in the conductance of the base emitter junction. Since the compensating diode is physically near the transistor there is a similar change in its conductance, and a compensating change is made in the base bias, keeping the collector-emitter current constant. The circuit of **fig. 2** is a very common integrated-circuit configuration.

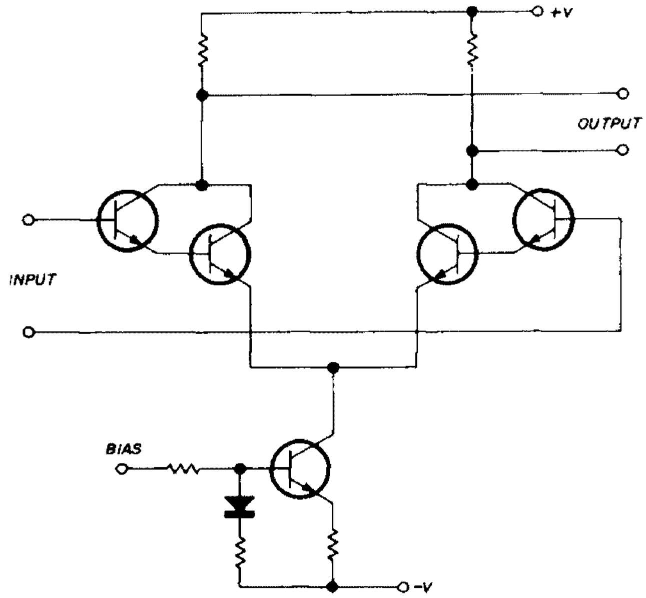


fig. 4. Differential amplifier with Darlington pairs.

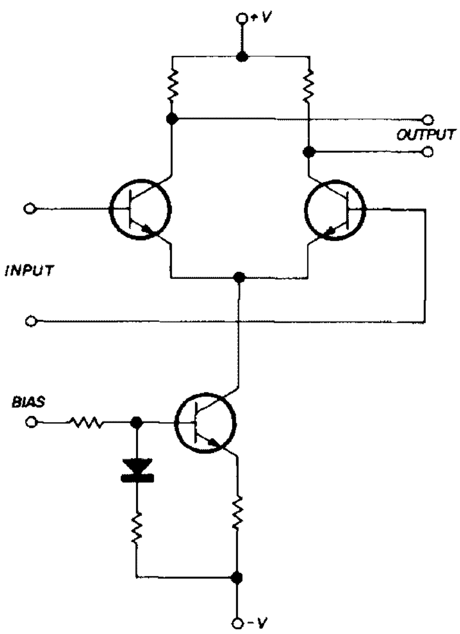


fig. 2. Constant-current bias source with temperature-compensating diode.

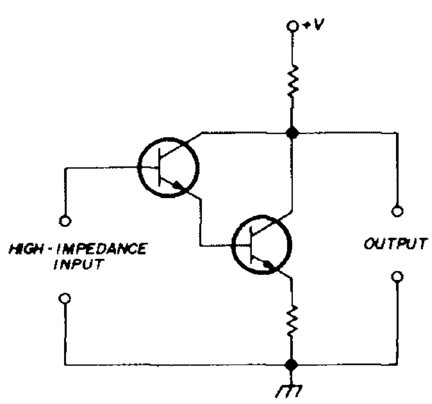


fig. 3. Basic Darlington pair.

darlington circuit

The differential amplifiers in **figs. 1** and **2** have low input impedances. High input impedances can be obtained by using Darlington circuitry which involves the addition of two more active elements. A simplified Darlington combination is shown in **fig. 3**; a typical application in an integrated circuit differential amplifier is shown in **fig. 4**.

In the normal transistor operation the base-emitter junction is forward biased and conducts. Resistance is low and approximates the product of beta times the emitter resistance. To some degree the input resistance can be increased by increasing the ohmic value of the emitter resistance at a sacrifice in gain. A better

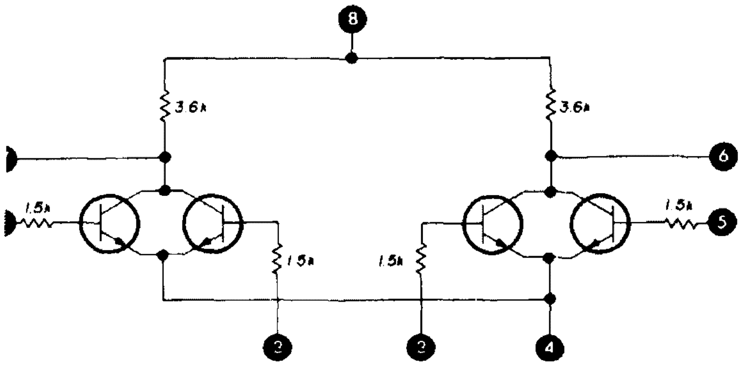


fig. 5. The Motorola HEP580 integrated circuit.

approach is to use the input resistance of a second transistor as the emitter resistance of the first transistor; the input stage then operates with a highly degenerative emitter circuit, and consequently, high input resistance. Both stages contribute output with a gain figure that is comparable to that obtained using a single transistor of the same type but operating with a much lower input resistance. Two such identical circuits are needed for the two separate inputs of a differential amplifier.

Motorola HEP580

The HEP580 is a low-cost integrated circuit composed of six resistors and four transistors. Internally the transistors are connected in pairs with separate base inputs (**fig. 5**). All emitters are joined together at pin 4. It is a basic differential-amplifier configuration using paired transistors instead of single devices. If desired, an external stabilizing constant current source can be added at pin 4; equal 3.6k collector load resistances are included. Series base resistances increase input resistance, reduce tendency to parasitic oscillations and provide additional isolation.

Fig. 6 shows two ways that are used to depict integrated circuits. The differential

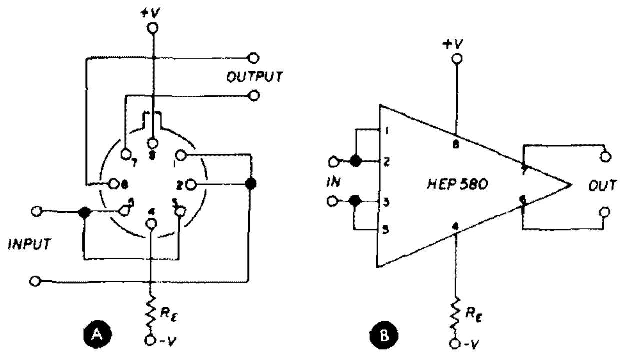


fig. 6. Two methods of representing the Motorola HEP580; arrangement in (B) is preferred.

circuit in **fig. 6A** is arranged around base pin designations of the ic. The triangular arrangement of **6B** is more common, and more instructive, because the circuit layout can be set down with well defined input and output sides regardless of pin numbers.

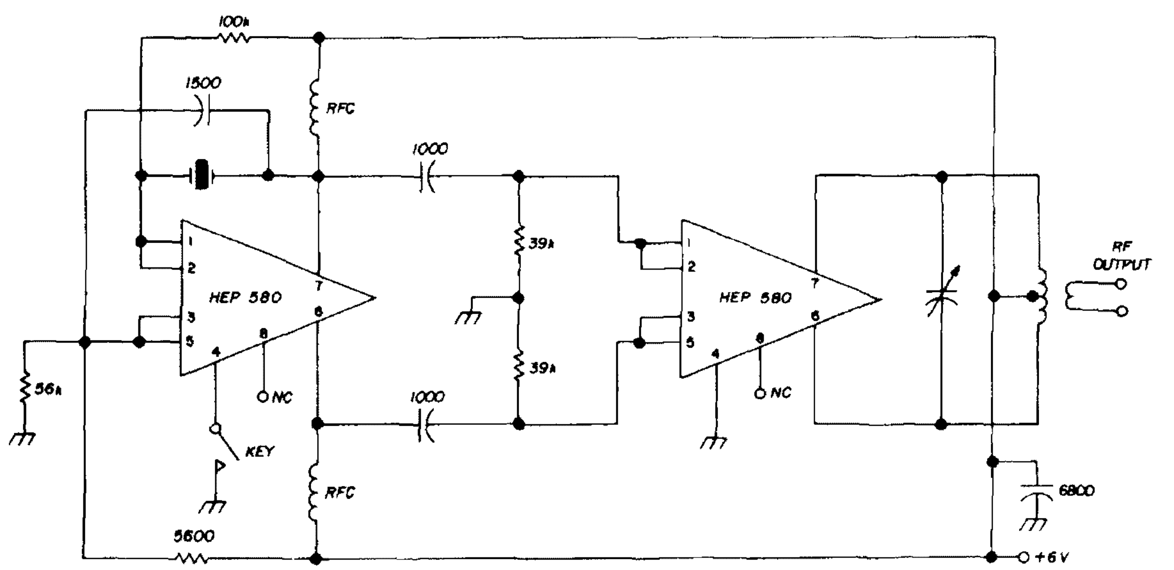


fig. 7. Integrated-circuit transmitter has output of 100 mW on 40 meters.

ic transmitter

A 100-milliwatt QRP transmitter can be built from two HEP580s, **fig. 7.** I have had no trouble working several hundred miles on 40 meters with this simple ic rig. The first section of one of the ics operates as a crystal oscillator; the second section as a phase inverter. Choke output is used, and approximately equal-amplitude and opposite-polarity rf signals are available for driving the output ic which operates as a push-pull amplifier. It will draw 20 to 30 milliamperes from a six-volt lantern battery. Dc input power is 120 milliwatts or more.

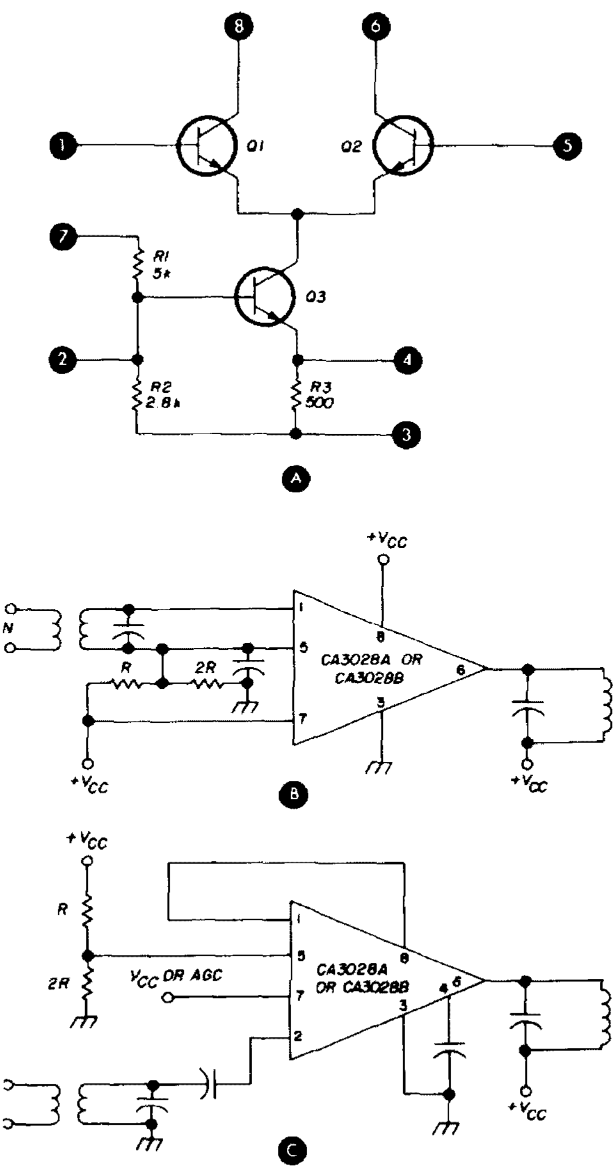


fig. 8. RCA CA3028 integrated circuit. internal circuit (A); as a balanced differential amplifier (B); as a cascode amplifier (C).

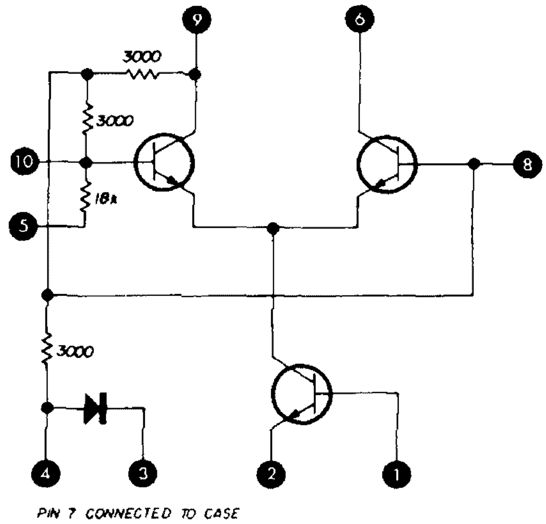


fig. 9. Circuit diagram of Motorola HEP590 integrated circuit.

RCA CA3028

The RCA CA3028 integrated circuit is a high-frequency unit that will function to 100 MHz and higher. It can be used successfully as an rf amplifier, converter, mixer, oscillator or limiter.

The internal diagram of the CA3028 is shown in **fig. 8.** The circuit is the classic *arrangement* consisting of a differential pair and constant-current bias source. Bias resistors are included. **Fig. 8B** shows the very few external components needed to use this ic as a differential rf amplifier. Signal is applied between pins 1 and 5 which connect to the bases of the differential amplifier. Output is taken from pin 6. Schematic **8C** shows how the same ic can be connected as a cascode rf amplifier. Signal is applied to pin 2 which connects to the base of transistor Q3. Its collector is direct coupled to the emitters of transistors Q1 and Q2 in cascode fashion. Output is taken from pin 6.

ZL4LV has used the RCA CA3028 integrated circuit as a balanced modulator, **fig. 10.**[1] The carrier signal is applied to the base of Q3 (pin 2) while audio is applied to the base of differential transistor Q1 (pin 1). The audio signal is applied in a differential mode while the

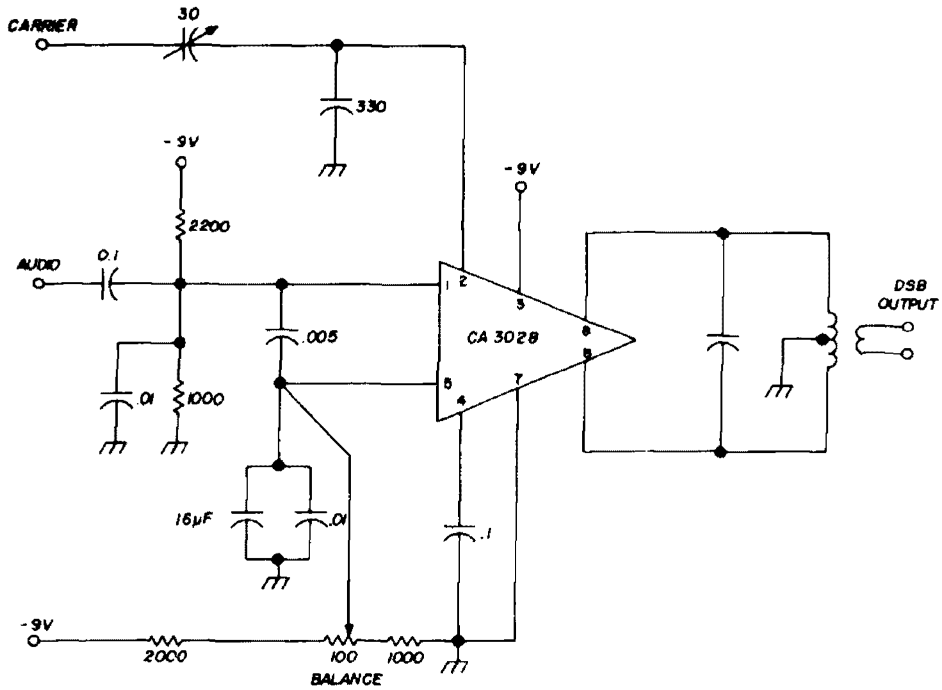


fig. 10. Integrated-circuit balanced modulator designed by ZL4LV.

carrier is applied as an in-phase component. Therefore, with proper balance, the carrier cancels in the collector-to-collector output circuit of the differential pair. Double sideband components are developed across the same output.

The Motorola HEP590 is a similar integrated circuit except that a temperature-compensating diode is a part of the package (fig. 9). Bill Hoisington, K1CLL,[2] has used the HEP590 successfully as an rf amplifier on both 6 and 40 meters, fig. 11.

Although these integrated circuits have been used principally in receivers they have dissipation ratings of several hundred milliwatts and would no doubt work well in QRPP transmitter circuits and in the earlier stages of QRP transmitters.

balanced modulator/demodulator

The Motorola MC1596G has been designed specifically for use in sideband systems. Internal circuit configuration and external circuit plan for a double-sideband suppressed carrier generator are given in fig. 12. Two differential amplifier pairs are included and incorporate individual transistors in their common emitter circuits to supply constant current bias. A second transistor is included in each leg for injecting the modulating signal. Carrier is applied in differential mode to the pairs of differential transistors. Outputs of the differential pairs are out-of-phase and under true balance the net carrier voltage is zero. Out-of-phase audio is applied to the transistors located in the emitter legs of the differential pairs. Upper and lower sideband frequencies develop across the output while the modulating wave is canceled.

The carrier signal is applied between pins 8 and 7; the modulating signal is between pins 1 and 4. Biasing for these latter two transistors is obtained from the -8 volt source connected to the arm of the carrier-null potentiometer. This

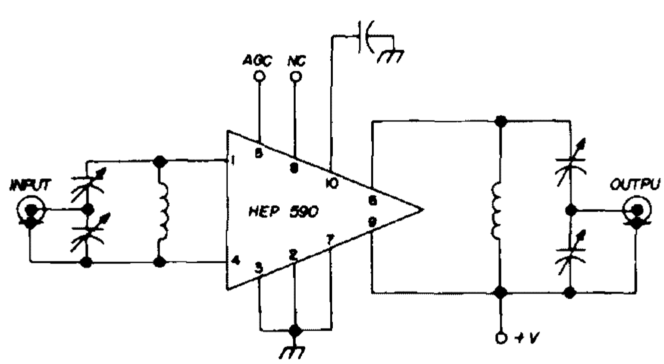
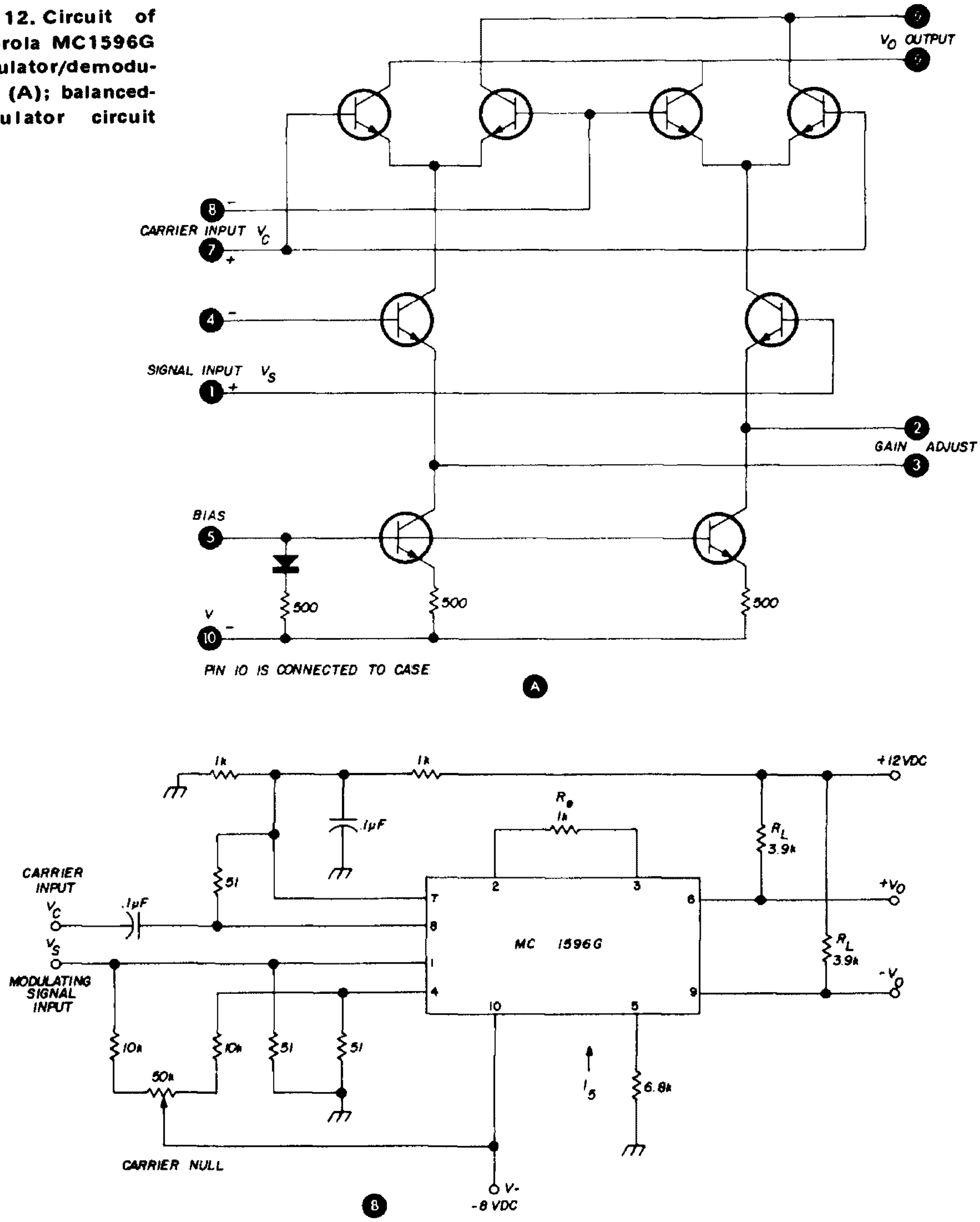


fig. 11. 50-MHz Integrated-circuit amplifier built by K1CLL.

fig. 12. Circuit of Motorola MC1596G modulator/demodulator (A); balanced-modulator circuit (B).



biasing sets bias and permits an appropriate adjustment for balancing out the carrier.

A balanced output is available between pins 6 and 9; single-ended output can be derived between either pin and common.

more power fets

The Siliconix 2N3970 is a switching fet that performs well as a high-frequency amplifier and oscillator. Its power output is about one-half that of a U222 power fet but at only one-quarter the cost. Device dissipation is 1.8 watt. Maximum drain voltage is 40, and in typical circuits the transistor draws 50 to 100 milliamperes. A TO-18 heat sink helps heat dissipation.

The 2N3970 performs well in a variety of oscillator circuits including the Miller, Pierce, Colpitts and push-pull. It oscillates efficiently 10 through 160 meters. A Pierce crystal oscillator and class-C amplifier is shown in fig. 13. This effective QRPP transmitter requires only a single resonant transformer.

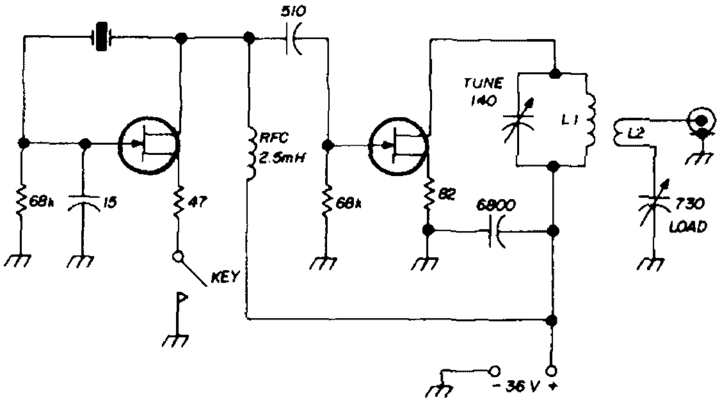


fig. 13. Simple two-stage fet transmitter for 40, 80 and 160 meters.

	160	80	40
L1	65 turns no. 26 on 1-1/4" coil form	40 turns no. 24 on 1-1/4" coil form	21 turns no. 22 on 1-1/4" coil form
L2	20 turns no. 26, bifilar wound on cold end of L1	13 turns no. 24, bifilar wound on cold end of L1	7 turns no. 22, bifilar wound on cold end of L1

Note how the fundamental class-C fet circuit closely matches conventional vacuum-tube practice. The resistor-capacitor combination at the fet gate develops the required cut-off bias. The source resistor, like the cathode resistor of a vacuum-tube amplifier, limits device current to a safe value when rf excitation is lost. A source current meter shows a dip when the drain is tuned through resonance. Likewise, the magnitude of the dip current rises as antenna coupling is increased.

Power outputs up to one-half watt are obtained on 40, 80 and 160 meters with somewhat less on 20 meters. A supply voltage of 36-volts is obtained by connecting three 12-volt lantern batteries in series. Drain current is typically 60 to 75 mA.

Outputs of 1 watt and higher can be obtained on 40, 80 and 160 meters using the push-pull circuit of fig. 14. The circuit arrangement is similar to that given for the U222 160-meter cw transmitter presented in the April issue of *ham radio.*[3]

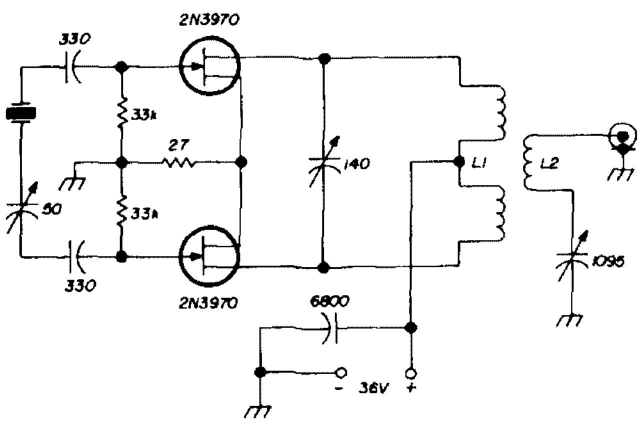


fig. 14. Push-pull fet power oscillator for 40, 80 and 160 meters.

	160	80	40
L1	65 turns no. 26 on 1-1/4" coil form, centertapped and divided	40 turns no. 24 on 1-1/4" coil form, centertapped and divided	21 turns no. 22 on 1-1/4" coil form, centertapped and divided
L2	20 turns no. 26 between halves of L1	13 turns no. 24 between halves of L1	7 turns no. 22 between halves of L1

References

1. Pat Hawker, G3VA, "ZL4LV High-Frequency Balanced Modulator," *Radio Communication,* July, 1969, page 466.
2. Bill Hoisington, K1CLL, "ICs for Amateur Use," *73,* October, 1970, page 22.
3. Ed Noll, W3FQJ, "Circuits and Techniques: Power FETs," *ham radio,* April, 1971, page 34.

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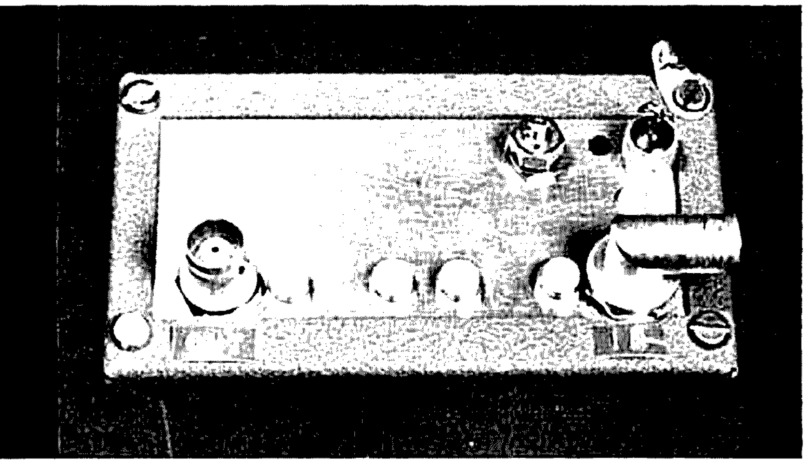


low-noise transistor 1296 MHz preamplifier

This high-performance 1296-MHz preamplifier provides a real breakthrough in noise figure and may spell the end for paramps on 1296 MHz

Dolph Vilardi, WA2VTR, 14 Oakwood Terrace, Spring Valley, New York

The state of the art in amateur receiving techniques has made dramatic strides during the last few years. Noise figures for devices available are now less than 1.5 dB at 432 MHz, and a recently announced Japanese transistor provides a noise figure of less than 3 dB at 1296 MHz with gain in excess of 13 dB.*

The 1296-MHz preamplifier developed by K2UYH,[1] and described in an improved two-stage version by myself[2] used KMC 5200 and 5500 transistors with noise figures less than 3 dB at 1000 MHz. Early reports of "around 3 or 4 dB at 1300 MHz" proved optimistic; noise figure measurements conducted by W2CCY, W2CQH and W2IMU showed that most devices were nearer to 5 or 6 dB with a few as low as 4.5 dB noise figure, although K2TKN claimed some devices he measured were around 4.0 dB.

With the best diode mixers available at that time the noise figures of the better front ends were measured optimistically at 7 dB – but most were nearer to 9 or 10 dB and many have been measured as high as 18 dB. With these noise figures it is easy to see why you needed two stages

*The Nippon Electric V766B, available in single units for $18 from California Eastern Laboratories, 87 Terrace Hall Avenue, Burlington, Massachusetts 01803.

to have sufficient gain to overcome mixer noise and establish a reasonable front-end noise figure.

inexpensive hot-carrier diodes such as the Hewlett-Packard HP2800. With such a converter it is now feasible to establish a

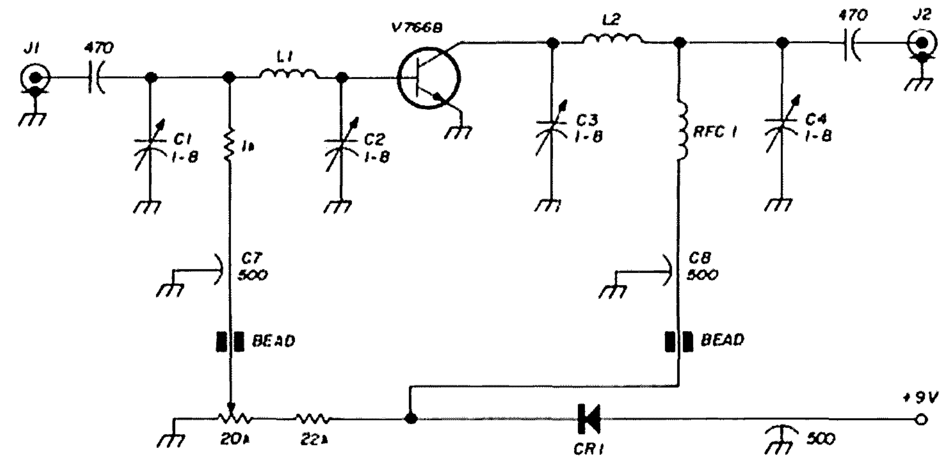
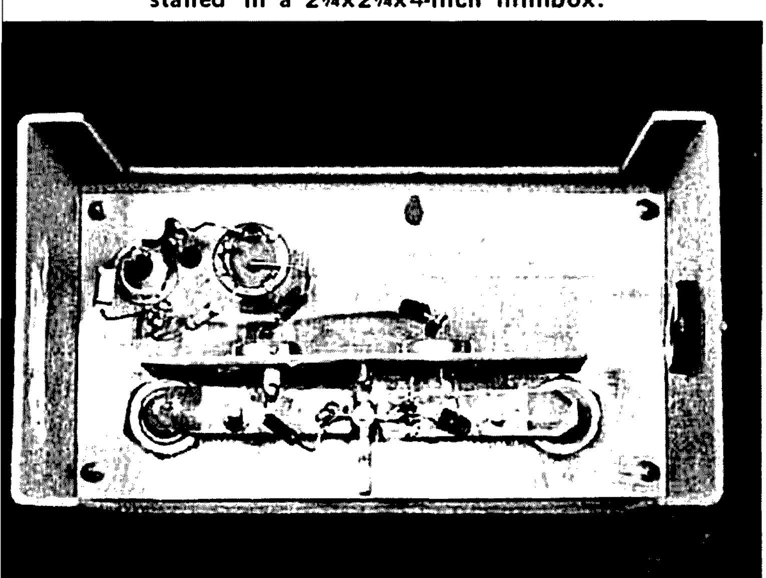


fig. 1. Circuit diagram and parts information for the low-noise 1296-MHz rf preamplifier. Construction is shown in fig. 2. Ferrite beads are available from Amidon Associates, 12033 Otsego Street, North Hollywood, California 91607; $2.00 per packet.

C1,C2,C3,C4	1 to 8 pF, high-quality short piston capacitor (Johanson used by author)
C5,C6	500-pF feedthrough, button mica or ceramic
CR1	protective diode, 10 mA or more
J1,J2	Teflon-insulated BNC chassis connector
L1,L2	brass strip, 3/8" wide (see fig. 2)
RFC1	2-3/4 turns no. 30 airwound on 1/16" diameter form

Modern converters with hot-carrier diode mixers and filters between the multiplier trough and the mixer[3] can achieve noise figures of about 8 dB with front-end noise figure based on the preamplifier parameters with a single stage *if* the preamplifier has a gain of at least 11 or 12 dB. The preamplifier described here meets these requirements.

The 1296-MHz preamplifier is installed in a 2¼x2¼x4-inch minibox.



the circuit

The circuit (fig. 1) is essentially that of the first stage of the 1296-MHz amplifier described in the 1970 ARRL Handbook with an rf choke substituted for one of the resistors. This provides slightly better gain. Also, the bias adjustment pot is changed to give smoother control with the parameters of the V766B transistor.

The physical changes from the original design are very important from the standpoint of stability and protection of the parts. Dimensions should be followed closely. An alternate and preferred method of mounting of the transistor is shown in fig. 3B. This construction makes it easy to remove and replace the transistor without damaging it.

construction

The preamplifier is built into a minibox for convenience and shielding, but all construction is done on 3/32 or 1/16-inch thick brass plate which is held in place by four screws (see fig. 2). This makes assembling and construction much easier as well as making the whole device very rigid. When working with the transistor, the collector lead is the longest one with the base lead opposite; the two remaining leads are emitter leads.

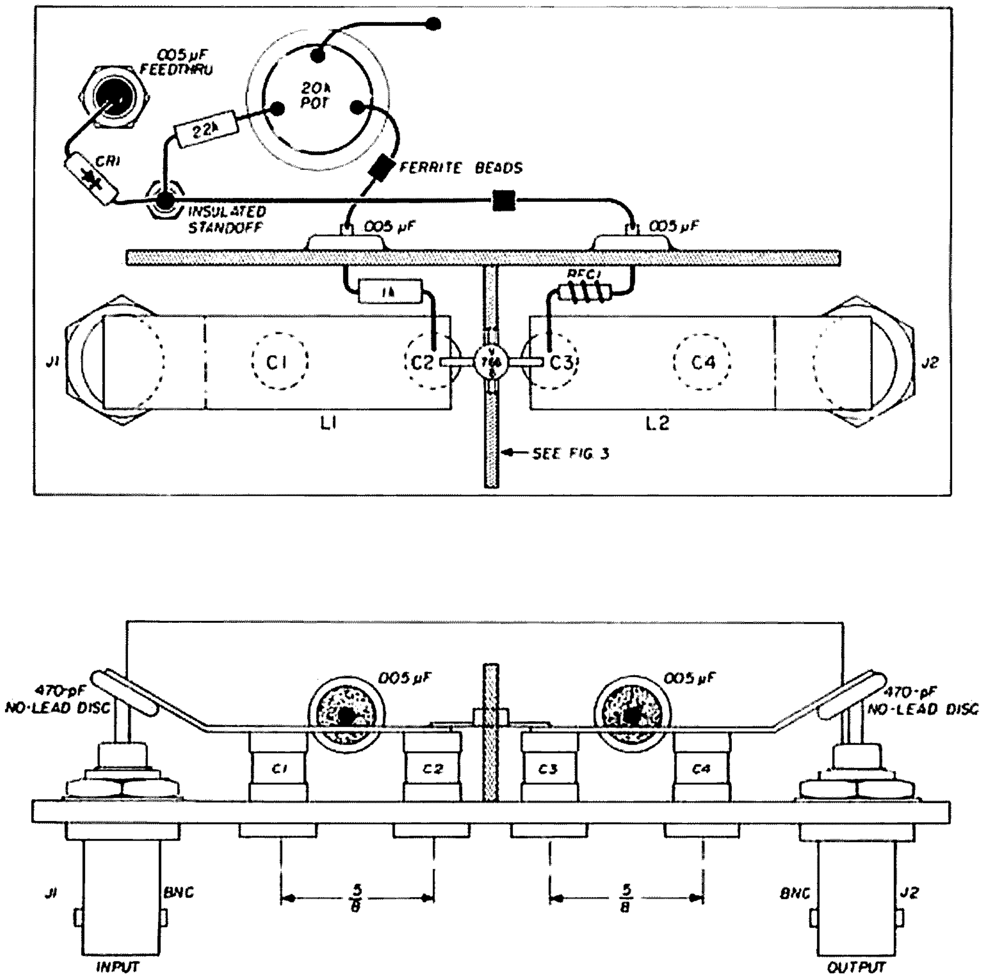


fig. 2. Construction details of the 1296-MHz preamplifier. For shield detail, see fig. 3. This unit is designed to fit into 4x2¼x2¼" minibox. RFC1 should be air supported; use a ¼-watt 1000- or 2000-ohm resistor if oscillations occur. This illustration is full size.

The minibox shown in the photographs is slightly deeper than necessary since I built an ac power supply into my unit. If the power supply is external the minibox can be much shallower.

If silver-impregnated epoxy is available it can be used at the transistor junctions and at the no-lead disc capacitors to avoid

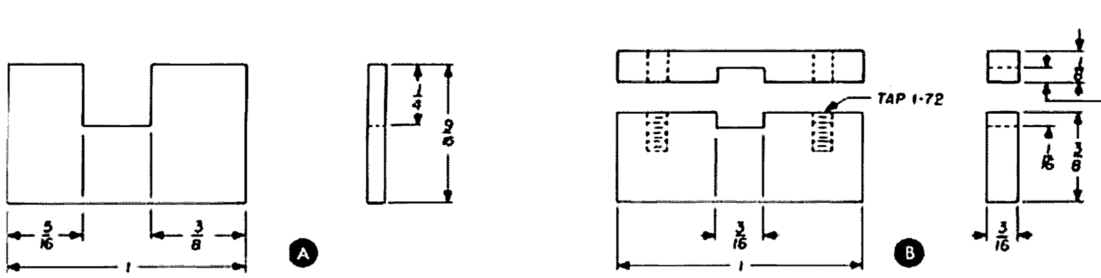


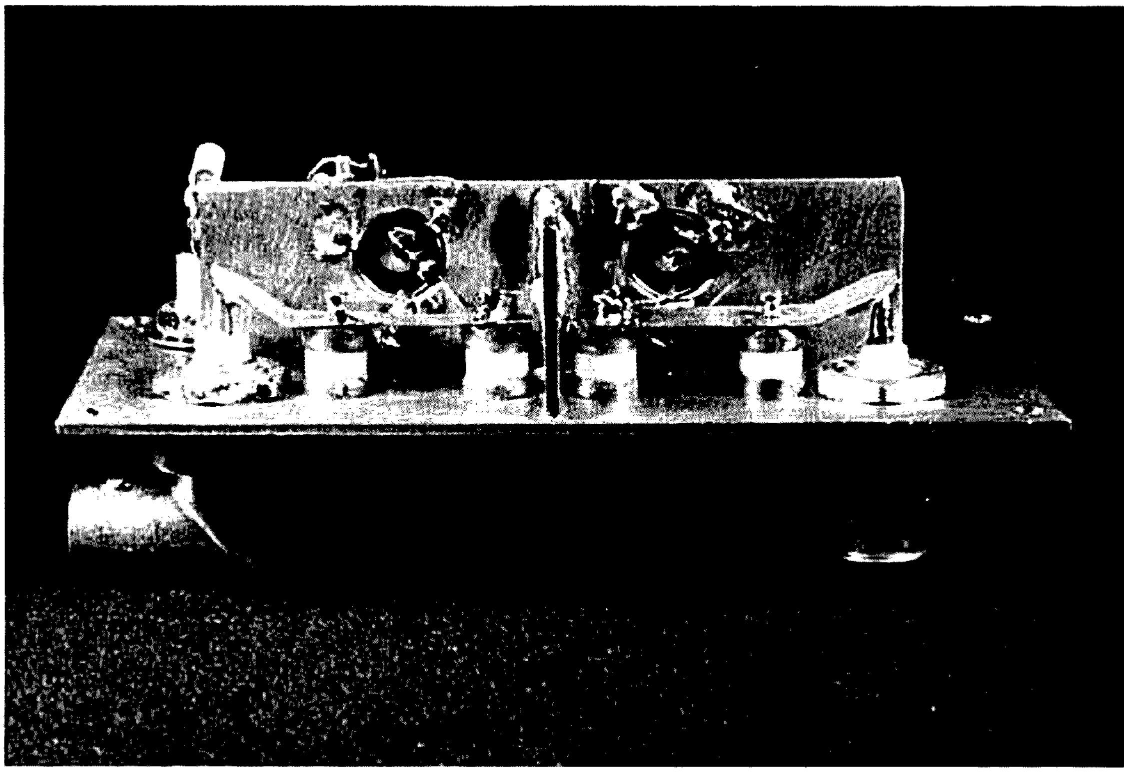
fig. 3. Transistor shield. Layout in (B) is preferred to arrangement in (A) as it permits easy transistor removal.

heat damage. The no-lead capacitors are difficult to find. You can make a good substitute by completely cutting off the leads from a ceramic disc capacitor and carefully filing the ceramic off the flat surfaces. You may spoil one or two disc capacitors but you'll eventually get the hang of it. Connections are soldered directly to the exposed surfaces.

The vertical partitions in the preamplifier are made of the same brass stock as the base and are preferably hard soldered, although regular "soft" solder will do.

Handle the transistor with care, especially when soldering it into the circuit. The Brookstone Company* sells a high-conductivity, low-temperature solder (TIX) which melts at 250° F and is excellent for this purpose.

The tuning capacitors are Johanson 0.8-8 pF units, but JFD equivalents, or any good quality *short* piston capacitors, will work. Use *thin* brass for the striplines to avoid overheating the capacitors when soldering. The brass striplines are mounted on top of the tuning capacitors and soldered directly to the tops. The capacitors are mounted on 5/8-inch centers; the part of the strap left over is bent upward at about 45° to accommodate the modified disc capacitors between the stripline and the BNC connectors.



Simple construction of the low-noise preamplifier for 1296 MHz.

*Available from the Brookstone Company, 5 Brookstone Building, Peterborough, New Hampshire 03458.

tuneup

Initial tuneup is best accomplished with a 1296-MHz signal (a typical 1296-MHz weak-signal source is shown in fig. 4). Apply about 6 volts to the transistor preamplifier and monitor collector current with a 10-mA meter. Adjust the bias control so collector current is 1 to 2 mA.

Start the tuneup procedure with all the capacitors at minimum capacitance; turn the output capacitors in one-half turn at a time until obtaining maximum reading on the receiver S-meter. Now adjust the input capacitor one turn at a time, repeaking the input capacitor near the transistor for maximum.

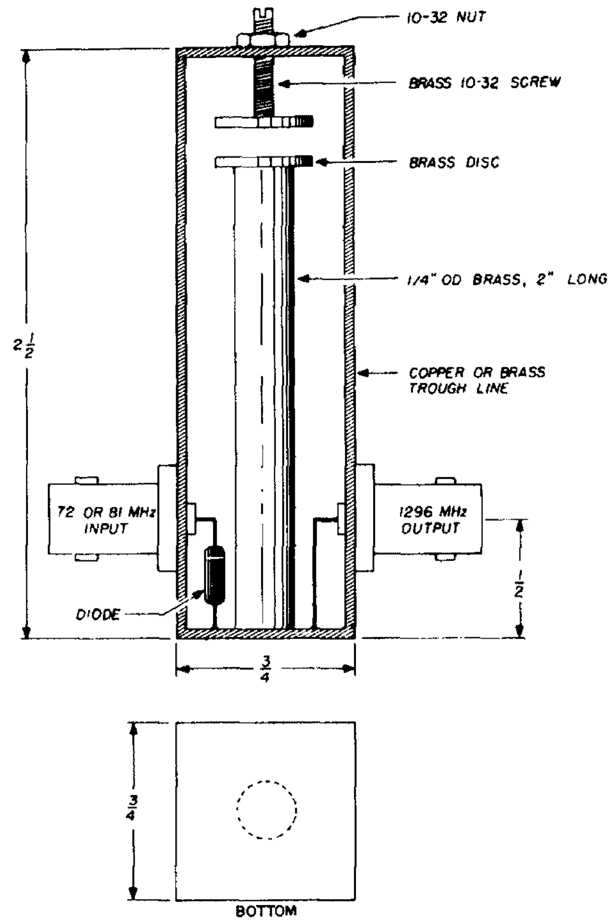


fig. 4. 1296-MHz weak-signal source uses 72-MHz injection and diode frequency multiplier. Diode may be a varactor, 1N914, 1N916 or 1N82. Input is link coupled to crystal-controlled oscillator.

Repeak all capacitors, and apply 9 volts to the preamplifier while adjusting collector current for maximum gain with lowest noise. *Do not exceed 6 mA collector current.* My V766B preamplifier worked best at 3.5 mA.

When tuning up (or using) the preamplifier do not allow the transistor to go into oscillation (as evidenced by sharply increased collector current) for more than a very short time or the transistor will be destroyed. The input and output networks are essentially pi networks but can be tuned to other modes. The function of these pi networks is to provide the transistor with a proper match; if the input of your converter is not close to 52 ohms you may need a 3-dB 52-ohm pad between the preamplifier and converter.

Final tuneup must be accomplished with the antenna connected to the input terminals. Put the 1296-MHz signal source near the antenna and connect the transmission line to the amplifier. Tune in the signal and repeat all adjustments for best signal-to-noise ratio.

If you happen to purchase a particularly "hot" V766B transistor you may have trouble with oscillations, although this is very rare. However, if you do have oscillation problems, replace the collector rf choke with a 1000- or 2000-ohm, 1/4-watt resistor; this will reduce the Q and gain of the stage.

summary

With this new low-noise transistor many serious 1296-MHz enthusiasts believe now is the time to discard the cranky parametric amplifier – the so-called advantages are hardly worth the added effort and complexity of the paramp. As W2IMU said recently, in relation to this 1296-MHz preamplifier, "We have entered a new era in EME for the amateur."

references

1. A. Katz, K2UYH, "A 1296-MHz Preamplifier – That Works!" *QST,* November, 1967, page 32.
2. D. Vilardi, WA2VTR, "Two-Stage Transistor Preamplifier for 1296 MHz," *QST,* December, 1968, page 40.
3. "The Radio Amateur's Handbook," ARRL, 1970, page 414.

ham radio

"Boy, you're in for the surprise of your life when you get out of here!"

determining power dissipation ratings of transistors

The power dissipation of a transistor depends upon the size and efficiency of the heat sink — here's how to determine practical power ratings

Jim Nyffeler, WN9CGW, RR 3, Bluffton, Indiana 46714

Have you ever ruined a transistor by operating it in excess of its rated temperature? Or have you ever bought a transistor with a large dissipation rating simply because you weren't sure whether a smaller device could withstand the power requirement? If you have, then this article is for you. I will discuss how to keep a transistor below its maximum temperature, and you will find that in some cases a device rated at 100 watts may be good for only a fraction of that much power.

safe operating area

The data sheets of most power transistors provide a safe-operating-area graph. Fig. 1 shows such a graph for an imaginary transistor capable of dissipating 100 watts. As shown on the graph, the maximum voltage which may safely be applied to this device is 40 volts; maximum current is 10 amps. The space enclosed by the black line represents the safe operating area – that is, the values of current and voltage at which the transistor may be operated without exceeding its maximum dissipation capabilities. Any value of voltage and current within the enclosed area may be safely applied to the transistor.

For example, the intersection of 4 amps and 20 volts lies within the line, indicating that the device may be used at these ratings without damage. On the

other hand, the intersection of 5 amps and 30 volts falls outside the enclosed area, indicating that the transistor, when operated at these values, will be generating more heat than it can dissipate, and will likely be destroyed.

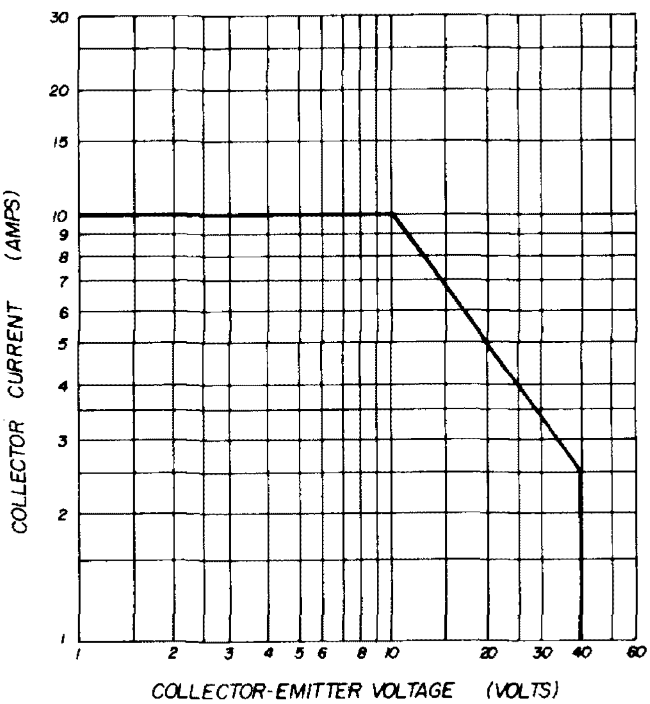


fig. 1. Typical safe-operating-area graph. The 100-watt transistor may be safely operated at any voltage and current values under the heavy line.

thermal resistance

The above explanation may seem very straight forward but unfortunately complications arise. Another factor, called *thermal resistance,* must be included in any calculations concerning a transistor's dissipation abilities. Thermal resistance is the resistance material offers to the conduction of heat. When used in conjunction with semiconductors, it is expressed in degrees centigrade per watt (°C/W). This means that for a specific number of watts dissipated, the temperature of the semiconductor will rise a definite number of degrees C above the ambient temperature.

It is possible to calculate the thermal resistance between the junction and the case (θ_{JC}) of any transistor if its power rating and its maximum permissible operating temperature, $T_{J(max)}$, are known.* These values can be found on the data sheet.

To find θJC for the 100-watt demonstration transistor, we will assume a metal-encased silicon unit. For such a transistor, $T_{J(max)}$ would be 200 °C. The thermal resistance may be found from the following formula:

$$\theta_{JC} = \frac{T_{J(max)} - T_A}{P_D} \qquad (1)$$

where T_A is the ambient temperature, P_D is the power rating of the transistor, and θ_{JC} and $T_{J(max)}$ are as explained above.

Inserting the values given for the 100-watt transistor at 25°C ambient:

$$\theta_{JC} = \frac{200 - 25}{100} = 1.75\ °C/W \qquad (2)$$

For a given temperature, the thermal resistance of the semiconductor determines the maximum power dissipation.

Before the transistor can be used, an additional thermal resistance, between the case and the ambient environment (θ_{CA}) must be known. This is because the heat generated at the semiconductor junction can only be transmitted through the case and must be dissipated into the surrounding environment whether it's air, a heat sink or a liquid coolant.

To determine θ_{CA}, you must know how the transistor is to be mounted. If the transistor is mounted in free, still air with no heat sink, θ_{CA} will be quite high and the power capabilities of the device will be greatly curtailed. The amount of power that may be dissipated with no heat sink depends on the junction-to-ambient thermal resistance, θ_{JA}. This factor can sometimes be found on the

*This value can also be found on the data sheet, but the reader should understand how it was obtained.

data sheet, and when used in **eq. 3** below it can be used to determine the maximum power that may be applied without a heat sink:

$$P_{D(max)} = \frac{T_{J(max)} - T_A}{\theta JA} \quad (3)$$

values for the 100-watt transistor:

$$\theta JA = 3^{\circ}C/W + 1.75^{\circ}CW + 0.4^{\circ}C/W = 6.15^{\circ}C/W \quad (5)$$

Hence, it is seen that the θJA of the complete assembly is 6.15°C/W. If this

table 1. Thermal resistance of typical heat-sink materials.

material	size (inches)	thermal resistance (°C/W)	remarks
finned heat sink	3x4x1½	3	bright aluminum
finned heat sink	3x5x2½	1.5	anodized aluminum
finned heat sink	5x5x4½	0.5	anodized aluminum
chassis	2x3	11.	unfinished aluminum
chassis	5x3	6.5	unfinished aluminum
chassis	7x5	4.5	unfinished aluminum
chassis	10x5	3.5	unfinished aluminum
chassis	12x10	2.5	unfinished aluminum
mica washer	—	0.4	with thermal compound

In the case of the 100-watt transistor used as an example, maximum dissipation would be approximately 5 watts.

However, if a heat sink is used the resistance to heat flow will be greatly reduced. Since heat sinks vary greatly in their ability to conduct heat, **table 1** has been included to give you an idea of typical values of thermal resistance. In addition to the heat sinks listed in the table, several chassis are also included since they are often used as sinks.

Assume that the first heat sink in the list, rated at 3° C/W thermal resistance, is to be used with the 100-watt transistor. To find the total thermal resistance between the semiconductor junction and the ambient, it is necessary to know the thermal resistances of the heat sink, the transistor and of any insulating washers used between the transistor and the sink:

$$\theta JA = \theta JC + \theta SA + \theta CS \quad (4)$$

The thermal resistance of a mica insulating washer is 0.4°C/W. Inserting the value is inserted into **eq. 3**, you can determine the maximum power that can be safely applied to the transistor:

$$P_{D(max)} = \frac{200 - 25}{6.15} = 34 \text{ watts} \quad (6)$$

In this arrangement the maximum power that can be dissipated by the transistor is only 29 watts. To safely apply more power to the device you must use a larger heat sink, or the sink must be cooled with forced air. If you try to dissipate more power than that calculated without additional cooling, the transistor will be destroyed.

You may ask why a semiconductor is rated at 100 watts when it cannot practically dissipate that much power. The answer is simple: Transistor manufacturers, having no idea of the specific type of heat sink to be used, publish the maximum power that the unit can dissipate when used with an *infinite* heat sink.

pulse operation

If a power pulse is momentarily ap-

plied to a transistor, higher dissipation is possible. **Fig. 2** shows a second safe-operating-area graph on which several dark lines are drawn. The line labeled *dc* is the maximum power-dissipation limit. If, for example, the demonstration transistor experiences stress for only 1 millisecond it

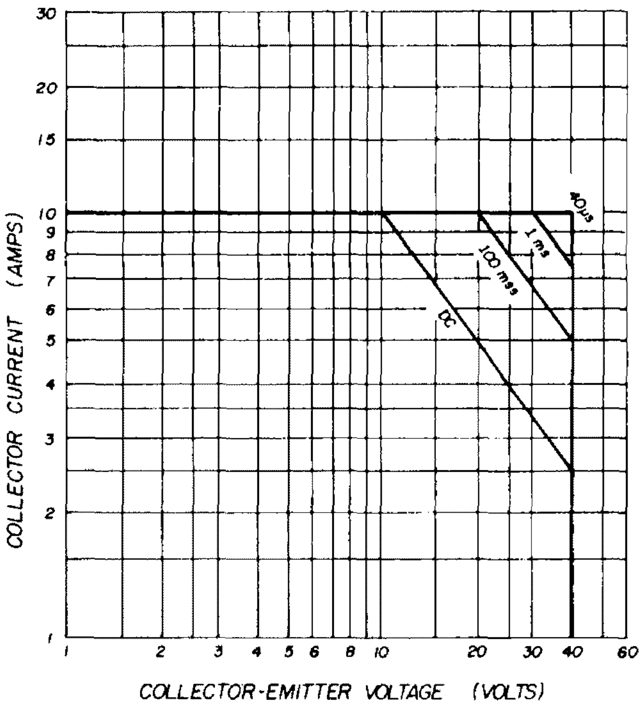


fig. 2. Safe-operating-area graph for single non-repetitive pulses.

can be seen from the graph that the device can dissipate 300 watts. Since the transistor is operated at three times its rated power the thermal resistance θ_{JC}, for a 1-millisecond pulse is reduced by a factor of 3 to 0.58°C/W. However, this holds true only if the transistor case is at a temperature of 25 °C prior to applying of the pulse. For other case temperatures the following formula is needed:

$$P_{D(pulse)} = \frac{T_{J(max)} - T_C}{\theta_{JC\,(pulse)}} \qquad (7)$$

where T_C is the case temperature and $\theta_{JC(pulse)}$ is the pulse thermal resistance as calculated above.

The case temperature may be found from:

$$T_C = P_{DC}\,(\theta_{CS} + \theta_{SA}) + T_A \qquad (8)$$

where P_{DC} is the steady state value of power being dissipated by the transistor prior to the application of the pulse, θ_{CS} is the thermal resistance of the insulating washer and θ_{SA} is the thermal resistance of the heat sink.

As an example, assume that the 100-watt transistor, using the heat sink discussed previously, is dissipating 10 watts just prior to the application of a 1-millisecond pulse. From **eq. 8** (at 25°C ambient)

$$T_C = 10(3 + 0.4) + 25 = 59°C \qquad (9)$$

Inserting the result into **eq. 7** gives:

$$P_{D(pulse)} = \frac{200 - 59}{0.58} = 243\ W$$

The transistor can withstand 243 watts for a period of 1 millisecond. Wattage ratings for other pulse widths are found in a similar manner by substituting the desired values.

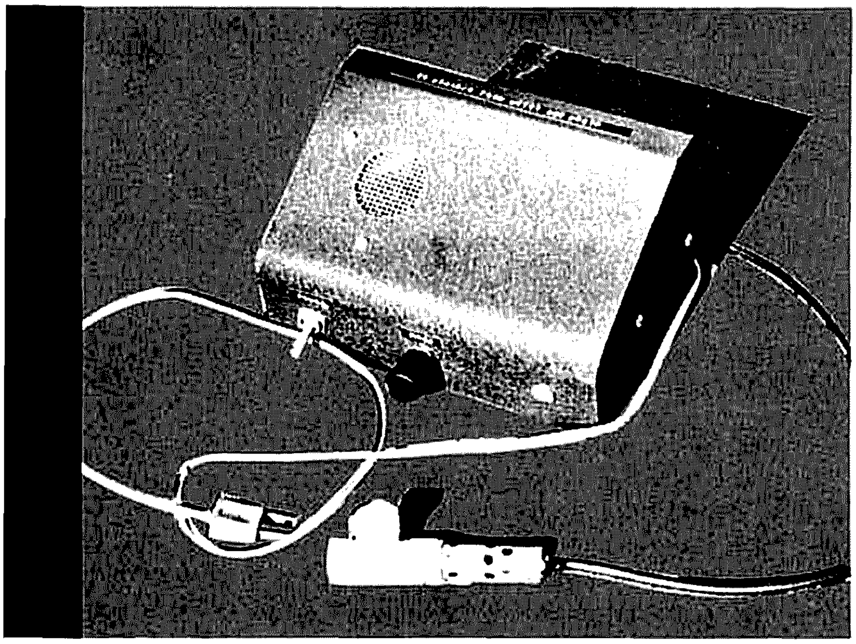
pulse-train dissipation

Since mathematical calculation of permissible power becomes rather unwieldy for repetitive pulses, at least one manufacturer, Motorola, has included graphs specifically for this purpose on his data sheets. As a result it is quite simple to find the maximum power dissipation for a wide range of pulse widths and for various duty cycles. A normalizing factor is obtained from the graph and multiplied by the dc value of θ_{JA}. The resultant factor may then be used in **eq. 3** to determine maximum power dissipation.

references

1. "The Semiconductor Data Book," 4th edition, 1969, Motorola Semiconductor Products, Inc., p. AN-16.
2. "RCA Power Circuits Manual," 1969, Radio Corporation of America, p. 71.
3. "RCA Transistor, Thyristor, and Diode Manual," 1969, Radio Corporation of America, p. 202.

ham radio



transmitter-tuning unit for the blind

The Noise Maker is an aural tuning meter that allows a blind person to tune his rig for maximum output

Don C. Miller, W9NTP, Waldron, Indiana 46182

Slow-scan television enthusiasts have a saying, "Hams should be seen as well as heard." I am an active slow scanner, but I was forced to admit one morning on 20 meters that this might not always apply, and in special circumstances, "Hams should be heard and not seen."

For many months several of us in the ninth call area have maintained contact with each other on the 20-meter band while driving to work on a newly built interstate highway. This highway provides superior radiation capabilities due to the excellent ground plane effect of the metal reinforcing in the concrete pavement. During this period of time W9TCT, myself and others were joined by a fixed station in the immediate area, K5MIB/9.

Wes soon told us that he was blind and that he lived in the area only during the winter months. Working Wes became routine, but his signal varied in strength from week to week and sometimes from day to day. It eventually occurred to us that this variation was due to the fact that his transmitter was not properly

tuned after moving from different parts of the band. When we inquired how he tuned his rig we were told that hams in the area occasionally dropped in and peaked the final.

On one particular morning Wes was so

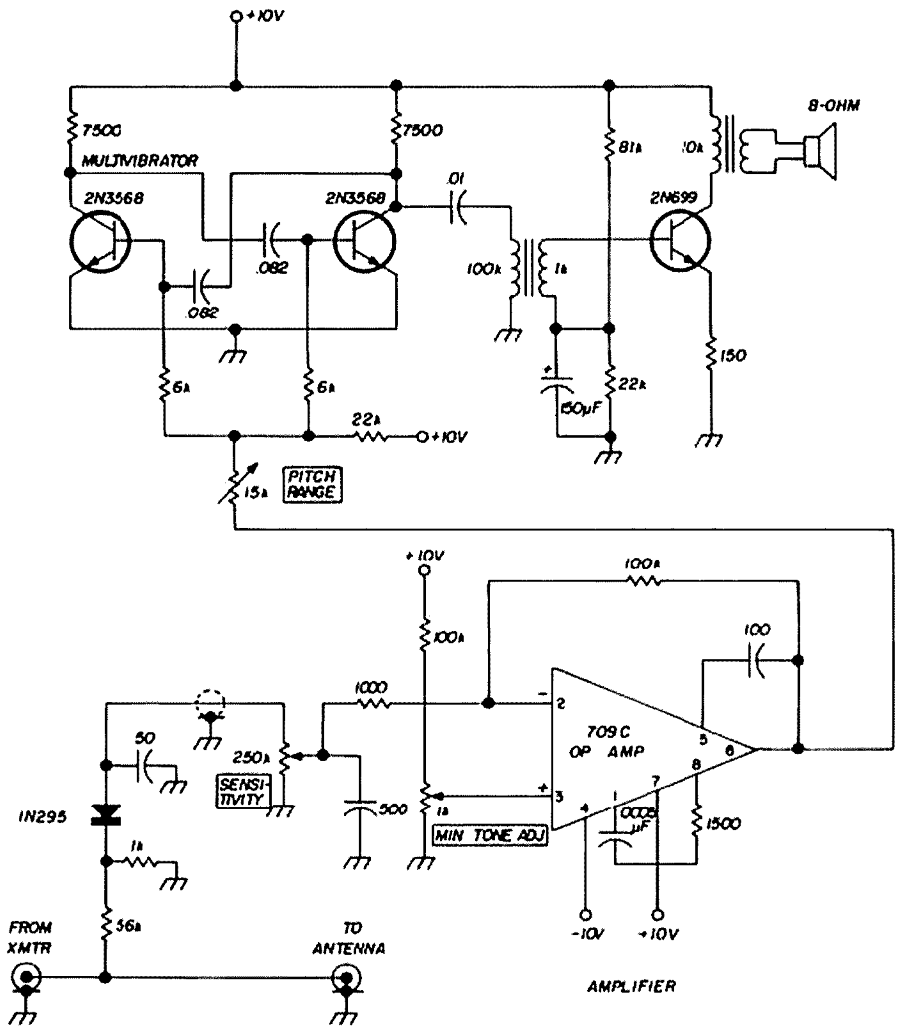


fig. 1. Schematic diagram of the aural transmitter-tuning unit.

Wes Bradley, K5MIB/9, tuning his Swan transceiver with the Noise Maker.

weak that his signal was barely readable just a few miles outside of town. W9TCT stopped in and tuned the rig for him. W9TCT and I felt that something had to be done, and decided that a simple device was needed to help Wes tune his rig to maximum output.

the circuit

Most slow scanners are familiar with the voltage-controlled multivibrator. The pitch of the oscillator can be changed over a rather wide range by providing a variable voltage to the base of the oscillator transistors. If the rectified rf current

from the output of the transmitter could be used to control the audio oscillator frequency a blind ham could easily tune his transmitter to maximum output by just listening to the highest pitch of the audio output frequency of the multivibrator. A simple diode output circuit and a 709C operational amplifier to supply the variable bias voltage were easily designed. A simple audio output transistor stage gave enough output to drive a small speaker.

theory of operation

The resistance divider consisting of the 56k and 1k resistors connected to the coaxial cable are adequate for power levels from 200 watts to 2 kW on all bands from 80 to 10 meters.

The gain of the 709C operational amplifier was set at 100. This amplifier has been compensated as recommended in the application sheets. A small voltage of about .01 volt is fed into the non-inverting input of the operational amplifier; this voltage provides the offset bias return of +1 volt for the base resistors of the multivibrator with no rf signal, telling the blind ham that the equipment is turned on.

The 709C amplifier has a positive swing of about +10 volts. The maximum pitch of the multivibrator can be changed over a small range by adjustment of the 15k pot. The lowest frequency is set by the small voltage fed into the non-inverting input of the 709C.

The multivibrator is conventional, and any npn transistors can be used. None of the components in this circuit are critical. The audio output stage uses a small imported high-impedance to low-impedance transformer to provide an impedance match between the collector of

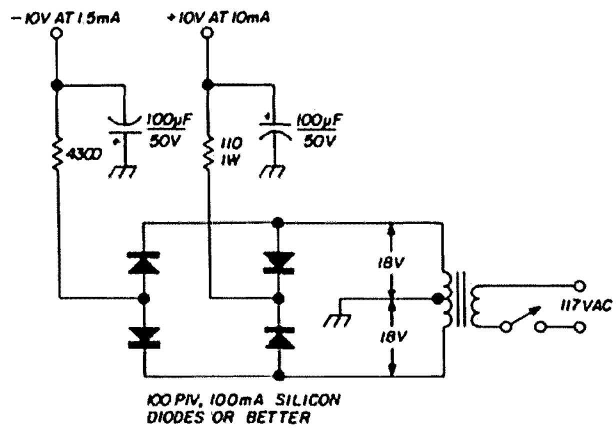
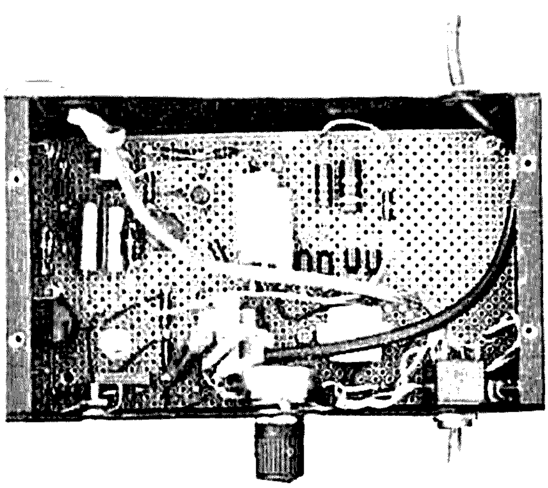


fig. 2. Power supply for the transmitter-tuning unit.



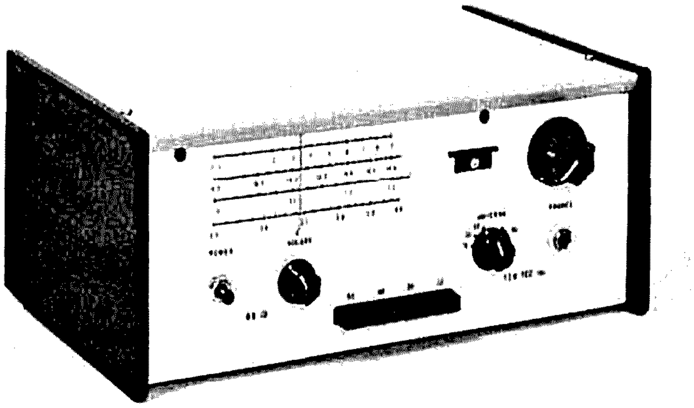
Internal layout of the aural tuning unit. Component layout is not critical.

the multivibrator and the base of the audio transistor stage. The output circuit consists of a small output transformer and speaker to provide the audible output signal.

A center-tapped junk-box transformer was used for the power supply. The current drain is extremely small, and available parts from around the shack should work nicely. The complete circuit is shown in fig. 1. An Amphenol T-connector was used to provide the housing for the rf detector circuit shown in fig. 1. The Noise Maker was enclosed in a small sloping metal cabinet. The entire circuit was built on vector board.

In the photograph K5MIB/9 is happily adjusting his Swan 350 to maximum output. Strange as it might seem, several hams that have sight have asked for the circuit for their rigs. I am indebted to W9TCT for help and inspiration in the design and construction of the Noise Maker.

ham radio



Ten-Tec RX10
communications receiver

A lot
of performance
in a small,
low-cost package

Skip Tenney, W1NLB

In these days of increasingly sophisticated and complex electronic communications equipment it is a pleasure to see a new product which fills an important need in a very simple and effective manner. The new Ten-Tec RX10 communications receiver is a natural for the beginner, or for the old timer who is looking for a good emergency receiver. The RX10 will give many hours of service from a handful of flashlight batteries.

In the Ten-Tec RX 10, direct conversion is used to provide complete coverage of the 80-, 40-, 20- and 15-meter amateur bands. The critical conversion stage uses an RCA 40604 dual-gate mosfet to provide low noise figure, high sensitivity and good overload characteristics. The balance of the receiver is built around straightforward bipolar devices. An oscillator/buffer/multiplier section provides an appropriate injection signal to the 40604 mixer, while a filter and four-stage audio amplifier arrangement deliver a more than adequate signal to the headphone output.

An added feature is the cw practice oscillator that is included; this helps the beginner get going with his code-practice program. The receiver also includes a built-in 115-volt ac power supply for home use.

The RX10 receiver is quite stable. The manufacturer claims no more than 100-Hz drift from turn on, and our experience at *ham radio* backs this up. Just for fun we connected the RX10 to a HAL Devices ST-6 RTTY Demodulator and tuned it to a teletype signal on 20 meters. Much to our surprise and delight we were treated to nearly ten minutes of perfect "hands-off" copy before the station signed. And this was with a cold receiver!

Although no specific measurements were made, the sensitivity was very good on both the 80- and 40-meter bands; both 20- and 15-meter performance seemed to

RX10 Specifications

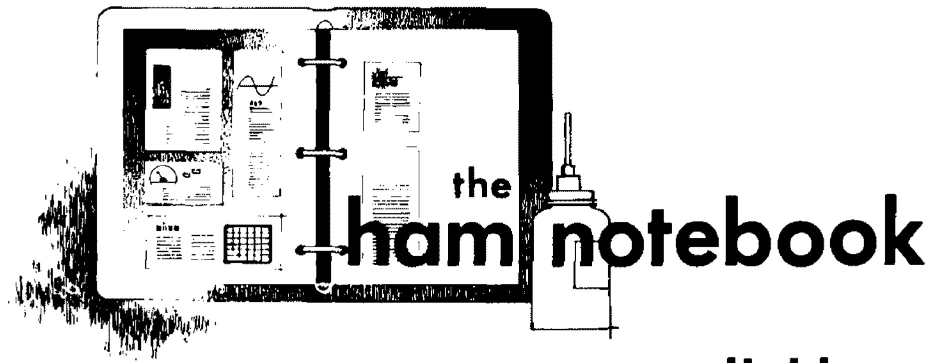
Frequency range	3.5-4.0 MHz, 7.0-7.3 MHz, 14.0-14.6 MHz, 21.0-21.9 MHz
modes	upper and lower ssb, cw and a-m
sensitivity	less than 1μV provides readable signal
stability	less than 100-Hz drift; no warm-up
selectivity	2 kHz at 6-dB down
audio	3 volts across 1000-ohm load
antenna	50 to 75 ohms, unbalanced
power	115 Vac, 50/60 Hz, 1/8 A, or 12 Vdc, 35 mA
size	10-3/8" wide, 4-1/2" high, 6-5/8" deep; 2¼ pounds
price	$59.95

offer more than satisfactory operation although it was a bit lower than on the lower bands. However, in all cases performance was well above the simple tube-type superhetrodynes which beginners have traditionally used.

Selectivity was also quite acceptable, even in the crowded 80-meter novice band. An audio filter with a 2-kHz bandwidth at 6 dB seemed to handle things quite well; in conjunction with the reasonably slow tuning rate for a simple receiver, it made it easy to find your way around. Of course, with the direct-conversion system you will hear the audio image of nearby stations, but in view of the excellent performance of this modestly priced receiver, I don't think this is an important consideration.

Please don't go out and buy a Ten-Tec RX10 for RTTY. However, I am sure that you will agree that for $59.95 there is a lot of performance in this little package. Ten-Tec products are available from your local dealer or Ten Tec, Inc., Sevierville, Tennessee 37862.

ham radio



economical decade standards

Resistor decades are convenient for experimental use but are bulky and expensive, especially the higher-accuracy units. A more convenient and compact precision, direct-reading variable resistance may be easily assembled using the ten-turn potentiometers and dials now available on the surplus market. Pots with accuracies of 1% and 0.1% linearity are available – which is more than adequate for most applications.

Potentiometer values in multiples of ten are used so that the selected resistance value may be read directly from the 10-turn dial by adding the proper number of zeros. Several potentiometer/dial units may be assembled in the form of the usual decade box.

The versatility of this arrangement may be expanded by bringing all three potentiometer terminals to binding posts on the front panel. With three binding posts continuously variable voltage dividers of accurately known ratio are conveniently available to the experimenter.

Gene Brizendine, W4ATE

switching counter readouts

One of the more expensive parts of a frequency counter, or counter dial*, is the indicating or readout equipment. As a result many applications provide the readout for only three or four digits of the count. In some cases you might want to read kHz but not individual Hz; in other cases you may want to read kHz or Hz but not MHz.

One way to obtain a readout when needed, without providing more Nixie tubes and associated storing and decoding circuitry, is to switch the indicating system from one part of the counter to another so *all* digits can be read when required. This can be accomplished by switching half as many indicators as the available total of digits counted, or by having just one digital indicator which is switched from one digit to another. A

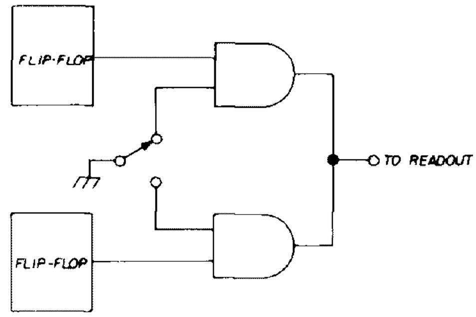


fig. 1. Using dual-input gates a readout can be switched to the Q output of either flip-flop to reduce cost. If the gates do not permit tying the outputs together, use steering diodes or a third gate.

*K. Macleish, W1EO, "A Frequency Counter for the Amateur Station," *QST,* October, 1970, page 15.

rotary switch, with four double-throw contacts per decade, will do the job.

Digital logic handbooks often show a "data selector" circuit using a gate in each data stream with provision for applying a signal to select which data stream

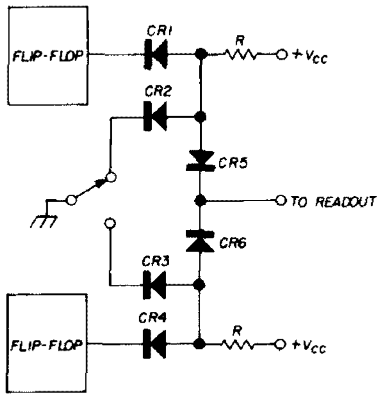


fig. 2. AND gates consisting of two diodes and a resistor can perform the switching between Q outputs, but require steering diodes CR5 and CR6.

appears at the output. **Fig. 1** shows one way to do this when the gate design permits output ORing. It uses a dual-input gate for each flip-flop output, or four gates per decade. Four more gates may be needed for the alternate decade which might be switched to the readout. A single toggle switch is sufficient for all gates.

Since no operation is required of the later flip-flops when the earlier ones are connected to the readout equipment, half the gates can be eliminated. This can be done by providing *preset* or *clear* signals or removing the B+ from the later flip-flops which are not to operate the digital readout indicator. If direct interconnection of flip-flop outputs causes a problem, steering diodes can be inserted.

Surplus diodes and a few resistors can perform the gate functions if a steering diode is included between the gate and the digital readout. This diode prevents the two interconnected gates from operating as a single gate. Again, a single toggle switch is sufficient for all gates involved in the readout switching (see **fig. 2**).

In addition to other types of diode switching that probably can be worked out to do the job at little cost, some other simplifications appear possible. One possibility is the elimination of one gate on the output of the tail-end flip-flops, turning these off by other means.

RTL and DTL flip-flops usually have their Q outputs connected to transistor collectors; the collectors are connected to $+V_{CC}$ through a resistor. Shorting the output of these *circuits may* overheat the resistor (although it *can* be replaced externally to the IC. Putting $+V_{CC}$ on the Q outputs can result in too much collector voltage on the transistors.

One simplification which I have not tried is shown in **fig. 3.** Diode CR1 forms a *one-input gate* when FF2, in an earlier position in the counter, is connected to the readout by the switch placing $+V_{CC}$ on the resistor. Whenever FF2 is *high* there is little voltage drop across R, so the *high* passes through steering diode CR3 to

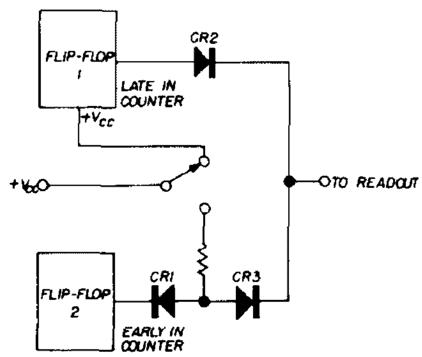


fig. 3. Simplified diode switching of readout when flip-flop 2 is earlier in the counter string than flip-flop 1.

the readout equipment. When FF2 is *low* the voltage from R goes into the flip-flop, creating a *low* input to the readout equipment. The other switch position puts $+V_{CC}$ on FF1 and other tail-end flip-flops to be read out. This allows these flip-flops to operate, feeding the FF1 Q output through steering diode CR2 to the readout equipment. Diode CR2 prevents the voltage across the resistors from feeding back into the Q output of FF1 when FF1 is turned off by the switch. CR2 may not be necessary if the voltage on the Q output is not harmful when FF1 has no $+V_{CC}$ supply.

Bill Conklin, K6KA

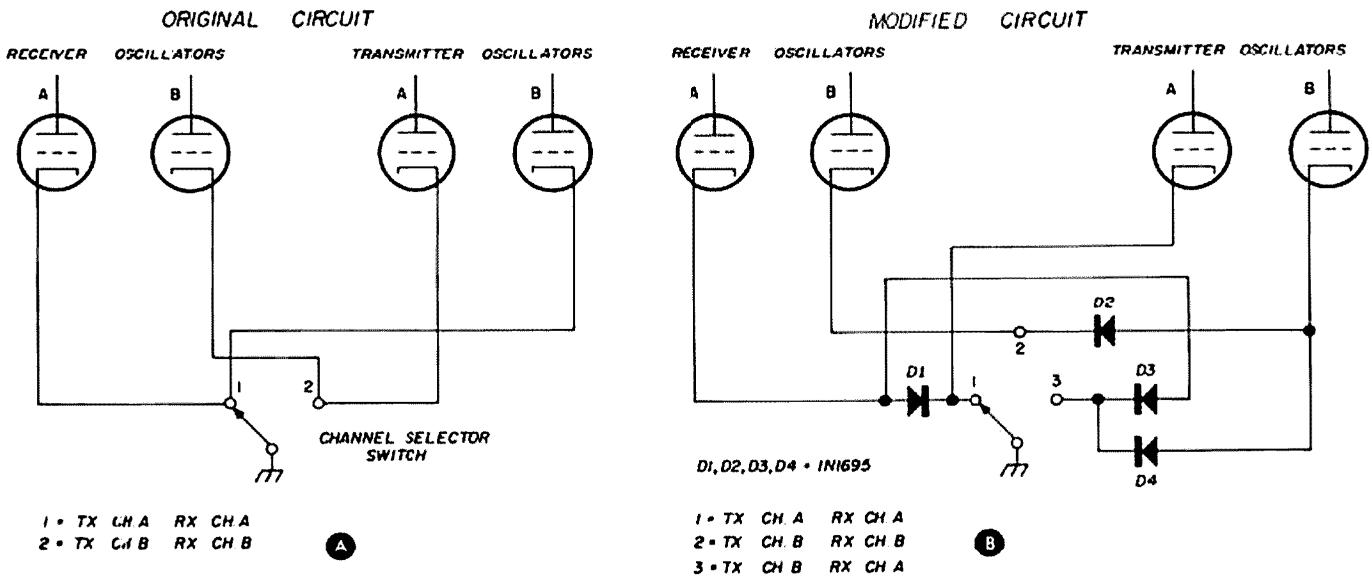


fig. 4. Method for increasing channel coverage in a typical fm oscillator circuit. Steering diodes, B, allow an extra transmit receive channel.

three channels from two

Many of the common fm mobiles are equipped for two-channel operation. However, with the different frequency combinations in repeaters using simplex, it might be necessary to transmit on channel B and receive on channel A (or vice versa). This can be accomplished easily with the addition of steering diodes (fig. 4). All you'll need are four diodes capable of carrying the current of the circuit and a single-pole, triple-throw switch. This system has been used in commercial applications for a private mobile system in this area.

The circuit shown in fig. 4B provides two simplex channels with a modified two-channel set. The idea for this application is from VE7BDY, to whom I'd like to express my thanks.

Vern Epp, VE7ABK

loose HW100 tuning knob

Many builders find that the main tuning knob on the HW100 is loose even though the dial mechanism is properly assembled and working well. The remedy is very simple.

Remove the tuning knob and place a washer on the end surface of the flexible spline. The washer must have a hole large enough to pass the collar on the spline on the vfo shaft (see fig. 5); otherwise the knob will not fit. The added washer fills up the space between the spline and the inside of the knob, applying an outward force on the knob that keeps it from wobbling. With just the right washer thickness, and an even coating of silicone grease on both sides, the action is "silky" smooth. If the washer is too thick, the knob will either not go on, or tuning will be very stiff. Several very thin washers may be needed to get the right feel.

Ai Lightstone, VE3EPY

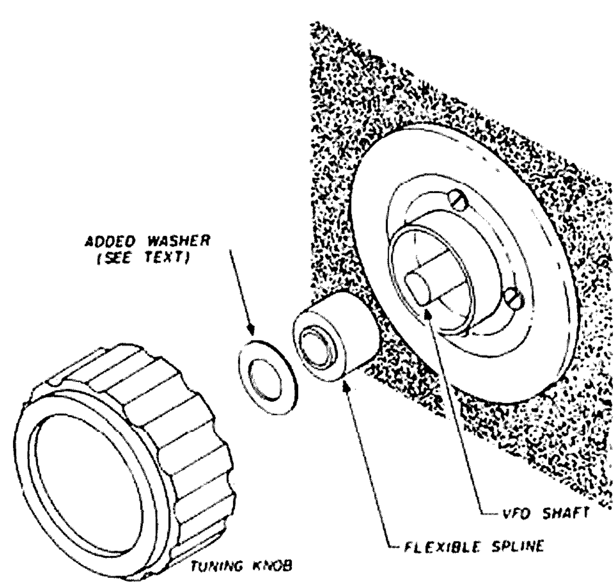
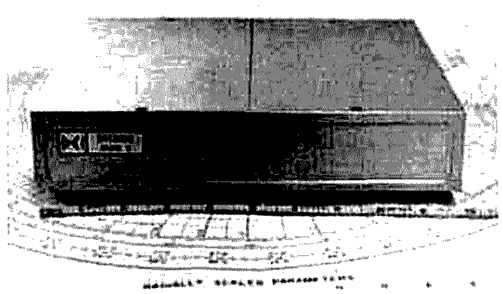


fig. 5. Remedy for loose tuning knob on the HW100.

new products

dycomm solid-state fm repeater



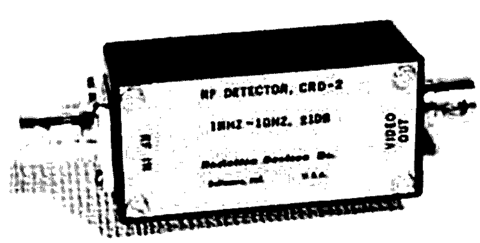
The new Dycomm *Echo II* is the first commercially available repeater for amateur radio use. It is completely solid state, uses no relays, is ultra stable, portable and FCC type accepted. The *Echo II* is designed to withstand the most severe environment, features 12 to 15 Vdc operation, multiple channel operation and comes equipped with full input/output option capability.

The *Echo II* transmitter will operate into an open or short circuit or any other mismatch. Input and output impedances are 50 ohms; connectors are type-N UG-58/U (mate with UG-1185/U). The unit has built-in carrier-operated relay, metering terminations for rf output, limiter current and discriminator output. The second optional frequencies may be remotely selected (by terminal connection) for either transmit, receive or both. Local/remote-control point is provided for local use as a transceiver so repeater may be used as a mobile *or* base-station transceiver. Guaranteed transmitter power output is 12 watts; typical output is 15 watts.

Useable *Echo II* receiver sensitivity is 0.2 μV; sensitivity for 20 dB quieting is typically 0.4 μV. Desensitization is less than 0.25 μV with 200-kHz channel spacing (no cavities). With 300-kHz channel spacing desensitization is negligible. Current consumption on receive, 40 mA; on transmit, 1.5 ampere. Inter-modulation interference, 70 dB minimum; spurious response attenuated more than 60 dB.

The repeater may be controlled with positive power up/down, timer, tone-burst entry, audio identification, etc. The *Echo II* is priced at $700 and includes two 6-dB antennas. For more information write to Dynamic Communications, Inc., Post Office Box 10116, Riviera Beach, Florida 33404, or use *check-off* on page 94.

rf detectors

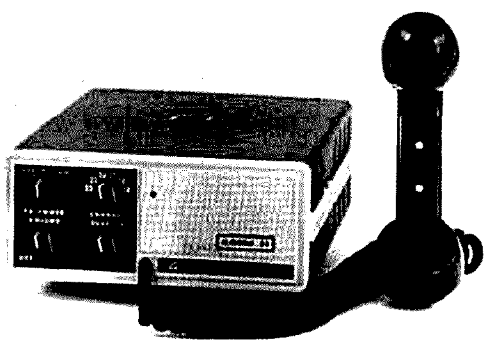


The new Radiation Devices CRD-2 coaxial detectors provide an economical means for rf demodulation and voltage measurement or monitoring over the frequency range from 1 to 1000 MHz. Units are available with or without terminating resistors. Either point-contact or hot-carrier diodes may be specified, with positive or negative output polarity. Frequency response is within 0.5 dB to 500 MHz; ±1.0 dB to 1000 MHz. Rectifica-

More Details? CHECK—OFF Page 94

tion efficiency is greater than 65%; maximum input voltage, 3 volts rms with point-contact diodes, or 25 volts rms with hot-carrier diodes. Lower frequency limit may be extended by addition of capacitor to internal terminals. Furnished with BNC or type-N input connector. Priced from $20 to $25, depending on options. For more information write to Radiation Devices Company, Post Office Box 8450, Baltimore, Maryland 21234, or use *check-off* on page 94.

two-meter fm transceiver



The new *Gladding 25* two-meter fm transceiver features 6-channels with 25-watts output. Designed and built by Pearce-Simpson, the world's largest manufacturer of marine communications equipment, the *Gladding 25* fills the requirement for the amateur who wants high power and multichannel fm capability. Crystals are factory-installed for 146.34/146.94 and 146.94/146.94 MHz.

The solid-state receiver uses an fet front end and integrated-circuit i-f strip. An 8-pole crystal filter provides superior selectivity. The transmitter is all solid state except for the driver and power output stages; power output is 25 watts in the *high* position, 1 watt in the *low* position. Output is ±5 kHz phase modulation with automatic deviation limiting. Harmonic and spurious emission is more than 60-dB down.

Receiver sensitivity of the *Gladding 25* is 0.5 μV. Squelch is adjustable, 0.4 μV or less for 80% rated audio output. Audio power output is 2 watts (10% distortion). Receiver spurious rejection is 60 dB or greater. Selectivity is ±7.5 kHz at 6 dB down; ±15 kHz maximum at 60-dB down.

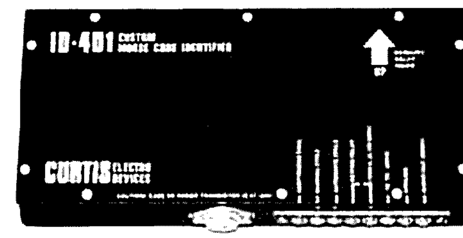
Nominal supply voltage is 13.6 volts dc, negative ground. Current drain on receive is 400 mA; standby, 1.2 amps; transmit, 10 amps. An optional matching ac power supply is available. The *Gladding 25* comes complete with mounting cradle and push-to-talk handset. Price, $249.95; accessory ac power supply, $69.95. Special combination price, transceiver and ac power supply, $299.95. For more information write to Pearce-Simpson, Division of Gladding Corporation, Post Office Box 800, Biscayne Annex, Miami, Florida 33152, or use *check-off* on page 94.

mobile antenna system

The Mosely *Rode-Master* is an all new amateur mobile antenna system that offers many significant money-saving options. The antenna covers 6, 10, 15, 40 and 75 or 80 meters with an adjustable vswr of 1.5:1 or better at any given frequency on each band. The *Rode-Master* features interchangeable coils for 10, 15, 20, 40 and 75/80 meters, and is power rated for 400 watts PEP ssb (200 watts a-m). The DX matching network is the reason Mosley can guarantee an adjustable vswr – the network is simple to install and operate the provides fine tuning on 20, 40 and 75/80 meters.

The upper mast section of the *Rode-Master* antenna doubles as a 6-meter whip, adjustable for the entire band. The telescoped whip-lock device permits precision tuning with little more than fingertip pressure. The antenna may be either bumper or trunk mounted, and includes break-over (hinge) for garaging or low overhangs. The antenna rotates 360° in the break-over position; this is convenient for coil insertion, antenna adjustments, etc. For more information, write to Mosley Electronics Inc., 4610 North Lindbergh Boulevard, Bridgeton, Missouri 63044, or use *check-off* on page 94.

fm repeater identifier



Curtis Electro Devices has announced a completely solid-state fm repeater identifier designed to provide low cost, reliable call-letter identification in accordance with FCC regulations. The new unit, the ID-401, provides an audio tone output in the form of a Morse code identification (such as DE W1DTY) in addition to carrier keying. Identification is transmitted initially on repeater activation and subsequently every three minutes as long as the repeater is being used. A final identification is transmitted after repeater activity ceases. Contact closure in the repeater control circuitry initiates identification.

Relay contacts are provided to key the repeater carrier when necessary. Provision is also made to transmit a continuous audio tone on command, and an audible monitor may be switched on for convenience in set-up. A cw output mode is available for direct keying of transmitters used in vhf aurora and meteor scatter work and similar applications requiring periodic identification.

The compact unit, which uses 12 complex integrated circuits, is completely enclosed in a heavy metal case and requires 700 mA from a −12 to −24 Vdc supply. Code speed, interval length and tone pitch and volume are adjustable. Operating temperature range is -40° F to $+140^\circ$ F.

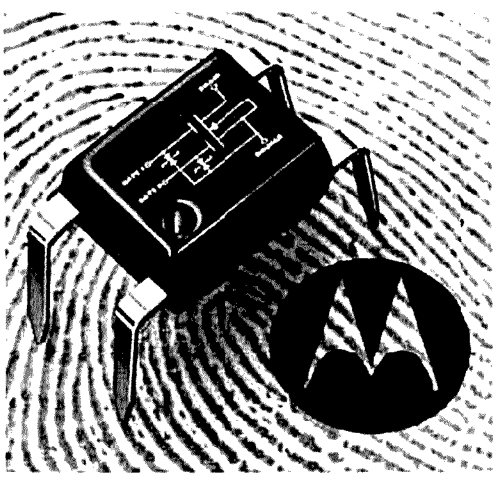
Code speed range of the ID-401 is 5 to 50 words per minute; interval time range is 1½ to 7 minutes. The capacity of the memory is 127 bits (one dot or one space = 1 bit). The unit has a built-in monitor speaker, and audio output of 0.5 volts p-p, 400 to 1500 Hz; the internal keying relay is rated at 2 amps, 500 volts,

More Details? CHECK—OFF Page 94

100 volt-amperes.

Price is $129.95, FOB Mountain View, California for the complete unit including a custom programmed memory. For more information, write Curtis Electro Devices, Box 4090, Mountain View, California 94040, or use *check-off* on page 94.

motorola mosfets

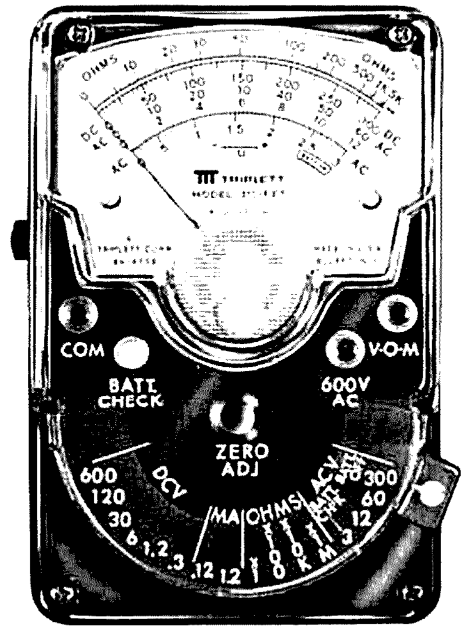


Three new dual-gate mosfets furnish low-cost high-performance amplifier/mixer applications in communications equipment, i-f strips and demodulators. The MPF120-122 are 50-cent range, plastic flat-pack cased devices with efficient agc control, low cross-modulation distortion, low feedback capacitance and high power gain plus gate diode protection. The MPF120 is an rf amplifier to 105 MHz with two separate channels. It provides excellent agc action, and a zener diode across the gate that shunts out voltage transients, adds reliability and stability.

The MPF121 is a vhf amplifier to 200 MHz. The MPF122 mixes rf with guaranteed frequencies of 104 and 244 MHz (optimum I_{DSS}). The new series of mosfets uses Motorola-developed silicon nitride passivation that ensures long-term stability under high-temperature and reverse bias conditions. Cross-mod for any of the devices is 1% (typical) with 100 mV of unwanted signal.

For more information, write to Motorola Semiconductor Products, Inc., Post Office Box 20912, Phoenix, Arizona 85036, or use *check-off* on page 94.

fet vom



The Triplett model 310-FET vom is an all solid-state instrument that features 10-megohm input impedance. The battery-operated unit has a single selector switch and provision for attaching an ac clamp-on adapter. It has high sensitivity of 300 mV dc full scale for transistor bias measurements, and resistance measurements to 500 megohms. Open-circuit voltage is 1.5 volts on all ranges; maximum short-circuit current is 30 mA. Accuracy is 3% of full scale on dc, 4% of full scale on ac, and 3% of dc arc on ohms. The meter has a zero-center mark for null measurements, and a polarity-reversing switch for dc and ohms. Sixteen ranges include dc current to 1.2 mA, dc voltage to 600 volts, ac voltage to 600 volts and ohms to 50 megohms (center scale). Price is $74.00 complete with leads, instruction manual and batteries from Triplett Corporation, Bluffton, Ohio 45817, or use *check-off* on page 94.

multi-tone repeater panel

Alpha Electronic Services has introduced a new, smaller 10-tone repeater control panel, the RCP-760. This unit is

More Details? CHECK—OFF Page 94

especially designed for two-way radio repeater or shared repeater systems, where the high reliability of solid-state circuitry is required at low cost. The RCP-760 will respond to the receipt of any of up to ten tone frequencies by keying and modulating the transmitter with regenerated received tone. The desired tone frequencies are determined by merely plugging in circuit board modules; up to twenty combinations can be achieved by the addition of the RCP-769-1 accessory panel.

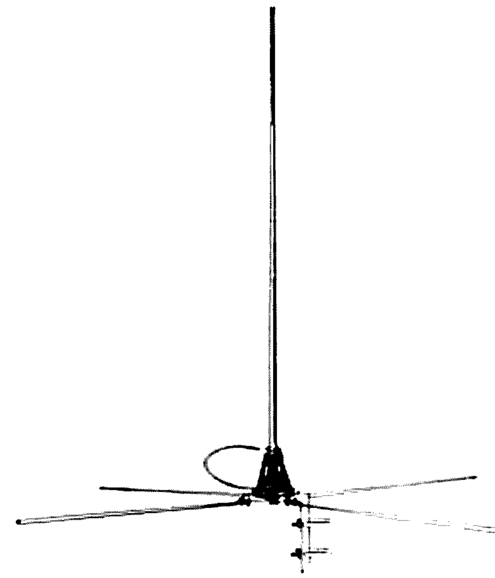
When ordered with at least one set of tone modules, the RCP-760 comes complete and ready to install. Built-in voltage regulation allows efficient operation over a wide range of input voltages (normal 12.6 Vdc). Dimensions are only 3-1/2 high, 3-3/4" deep, on a 19" panel. A 117-Vac Power supply is available.

For more information, write to Alpha Electronic Services Inc., 8431 Monroe Avenue, Stanton, California 90680, or use *check-off* on page 94.

communications ic

National Semiconductor has started production of an integrated circuit which may qualify as a general-purpose communications subsystem. The LM373, designed for a-m, fm and ssb applications, contains two amplifier sections (four gain/limiters), a gain-control stage, fully balanced fm and ssb detector, and an active a-m/ssb peak detector whose output matches the agc input characteristics. The bandpass characteristics can be shaped from audio to 15 MHz with a single external filter — crystal, ceramic, mechanical of LC. An LC tuned quadrature circuit gives 80-mV audio output for 75 kHz deviation at 10.7 MHz in a typical wideband fm application. In a-m operation typical sensitivity is 5 microvolts for 10 dB signal and noise. Price is $4.85 each in small quantities. For sales information, write to National Semiconductor Corporation, 2975 San Ysidro Way, Santa Clara, California 95051.

mosley two-meter antenna



The new omni-direction vertically-polarized *Diplomat 2* antenna from Mosley features low angle of radiation for maximum coverage. Gain is specified at 3.4 dB as compared to a ¼-wave ground plane, and vswr is 1.5:1 or better. Input impedance is 52 ohms. The vertical element is made from high tensile strength aluminum; the base is high-impact polystyrene. The *Diplomat 2* is rated at 1 kW a-m or cw, 2 kW PEP ssb (input to the final amplifier). The base mounting fits up to a 1½-inch mast. The *Diplomat 2* is available at your dealer for $10.58; similar antennas for six and ten meters, the *Diplomat 6* and *Diplomat 10,* are $21.75 and $29.30 respectively. For more information, use *check-off* on page 94, or write to Mosley Electronics, Inc., 4610 North Lindbergh Boulevard, Bridgeton, Missouri 63042.

new fcc form

Effective July 1, 1971, the Federal Communications Commission will accept for filing *only* editions of FCC Form 610, *Application for Individual Amateur Radio Station and/or Operator License,* dated July, 1970 or later. These forms are currently available. *No* applications on FCC 610 forms dated before July, 1970 will be accepted for filing after July 1, 1971.

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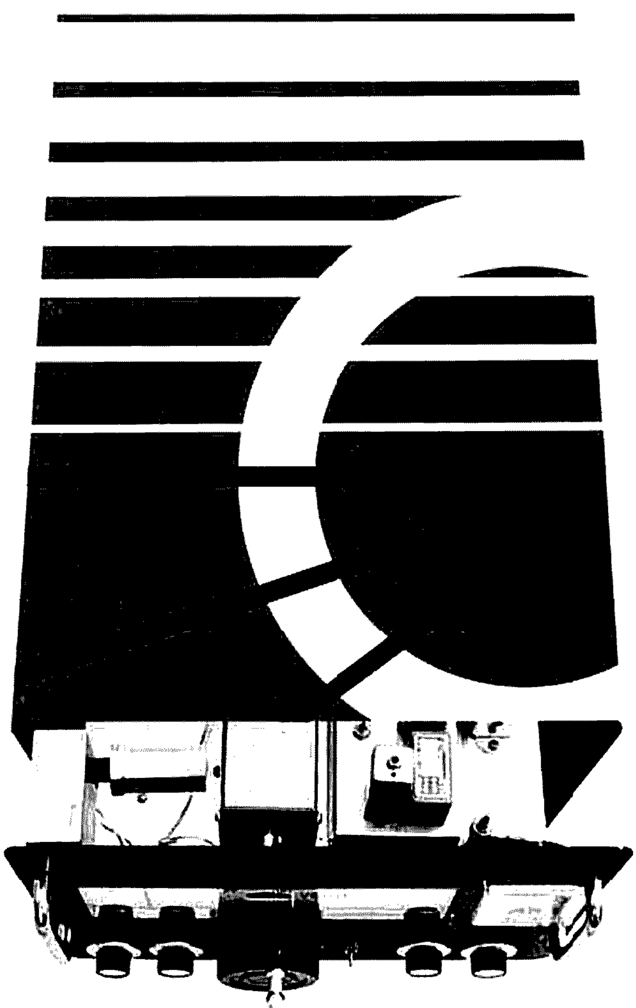
ham radio magazine

JULY 1971



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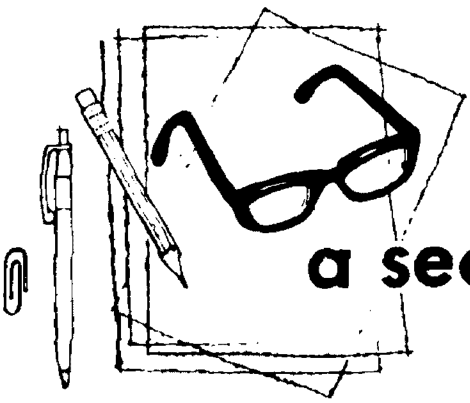


a second look

by jim fisk

Radio signals from a spacecraft on the other side of the sun have been used to prove, once again, the correctness of Einstein's General Theory of Relativity. This theory, which is an extension of the well known equations that relate mass and energy, is a geometrical theory of gravity, and is of tremendous importance to understanding the universe.

According to the General Theory of Relativity, electromagnetic waves near massive bodies, such as the sun, travel more slowly than they do in free space. Until recently, there have been three methods of testing this theory, all proposed by Einstein. One test involves the slowly-rotating elliptical orbit of Mercury, the planet nearest the sun. Einstein's equations predict the slight rotation of the orbit much more closely than the older gravitational theory of Sir Isaac Newton.

A second test predicts a slight change in the color of the light from the sun and the stars. The stars' intense gravitational fields cause the light they produce to shift toward the red end of the spectrum, as seen from the earth. Although this effect has been observed, it has not been measured accurately because the shift is very slight and difficult to measure.

A third test considers the apparent bending of starlight as it passes through the gravitational field of the sun. However, this can only be measured during a total eclipse of the sun so distant stars can be observed optically. When this rare alignment of the stars took place in 1919 Einstein's predictions were verified within the accuracy of the measuring equipment.

A chance for a fourth test of the General Theory of Relativity occurred recently when the Mariner 6 spacecraft passed near the sun. This is the same craft which flew by Mars in 1969 and measured surface temperatures and atmospheric composition, and sent back television pictures of the planet's surface. Since Mariner 6 passed very close to the sun's intense gravitational field, according to Einstein's equations the radio signals from the spacecraft would be slowed slightly.

To measure this signal delay radio signals were transmitted from earth to a repeater on board the Mariner 6 which returned the signals to earth. The normal round-trip time for the signals, not counting any gravitational affects, is estimated at 45 minutes. According to the General Theory of Relativity the radio signals would be slowed by 200 microseconds as they traveled through the sun's gravitational field.

Information recently released by the Jet Propulsion Laboratory in Pasadena, California indicates that the Mariner 6 radio signals exhibited a maximum relativistic signal delay of about 204 microseconds, within two per cent of that predicted by Einstein's Theory. If the sixteenth-century theory of Isaac Newton were correct there would have been no time delay at all.

Jim Fisk, W1DTY
editor

integrated-circuit communications receiver for 80 meters

A high-performance integrated-circuit ssb receiver featuring half-microvolt sensitivity, crystal-filter selectivity 100-dB agc range and low-drift vfo

Paul Hrivnak, VE3ELP, 16 Chatterton Boulevard, Scarborough, Ontario, Canada

Integrated circuits are being used more and more in the design and construction of amateur radio equipment. The idea for this receiver was originally conceived after reading K7WQR's excellent article on the Motorola MC1596G balanced modulator.[1] In the receiver described here the MC1596G is used as a double-balanced mixer.

The first receiver I built around the MC1596G is shown in block form in fig. 1. This original prototype had no front end, i-f amplifier or agc circuits. However, with the high conversion gain provided by the MC1596G* signals on 80 meters were copied with little effort.

The second design, shown in fig. 2, added an RCA 40673 zener-protected dual-gate mosfet rf amplifier stage ahead of the MC1596G mixer. However, manual control of rf and audio gain was still required to listen comfortably to signals on 80 meters. (The front panel has an rf gain control which is not used in the final design; this was used in the second prototype.)

An agc system for a receiver with no i-f stages presented a real problem. I was stuck until I discovered the Plessey SL-600 series of integrated circuits. One of these ICs, the SL621, is designed specifically for gain control circuitry; another, the SL610, is designed for rf amplifier applications. With the discovery

*Gain of the MC1596G is typically 24 dB when impedances are properly matched.

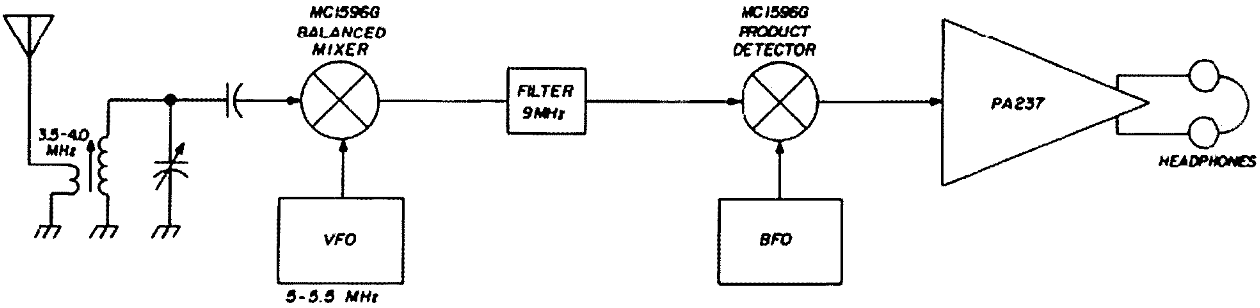


fig. 1. Block diagram of the original receiver design. No front-end, i-f strip or agc was required for copying signals on 80 meters.

of these vital links, the receiver design finalized into the form shown in fig. 3.

As can be seen from the schematic and photographs, modular construction was used throughout the receiver. This speeds construction time, allows bench testing of all modules before deciding on final mechanical design, and allows for circuit changes without upsetting the physical layout.

rf amplifier

The Plessey SL610C is a low-noise, low-distortion rf voltage amplifier with integral supply-line decoupling and provisions for agc. It has a 50-dB agc range with maximum signal handling capability of 250 mV rms.

The rf amplifier stage was built on single-sided copper-clad printed-circuit board. Generous quantities of Teflon

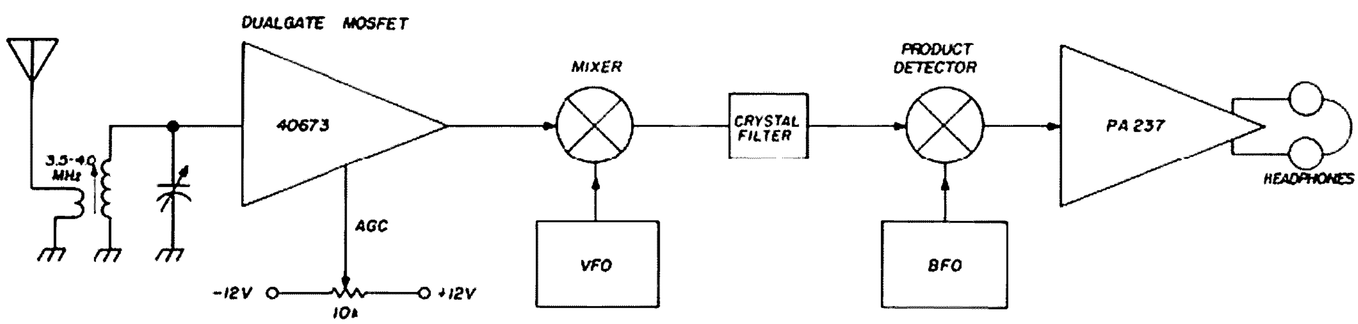
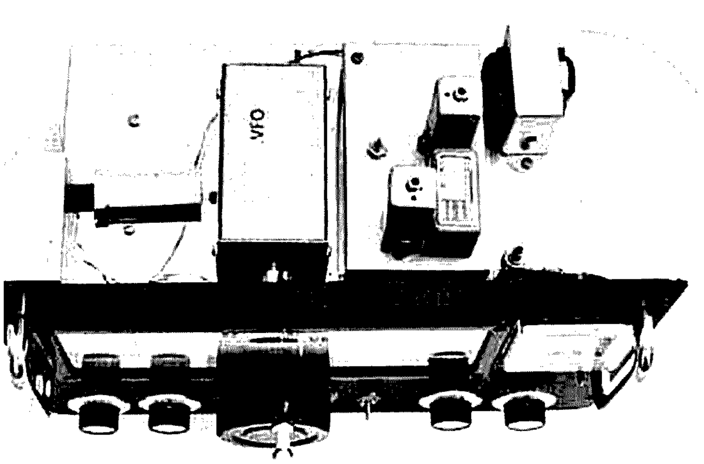


fig. 2. Second receiver design added a dual-gate mosfet rf amplifier stage.

Top view of the receiver with 100-kHz crystal oscillator on the left, 5-MHz vfo in center, and high-frequency mixer, crystal filter and power transformer to the right.



push-through terminals keep construction simple and leads short and direct. Miniature RG-174/U coaxial cable is used in the rf amplifier and to connect the modules together throughout the receiver. Expensive coaxial connectors are eliminated by running the coax through rubber grommets in the chassis.

The rf amplifier, as well as the other individual circuit modules, is housed in a small minibox; power-supply voltage is fed in through a 0.001-μF feedthrough capacitor. With this construction technique no oscillations were experienced; gain is approximately 18 dB.

mixer and filter

Except for the addition of ssb input

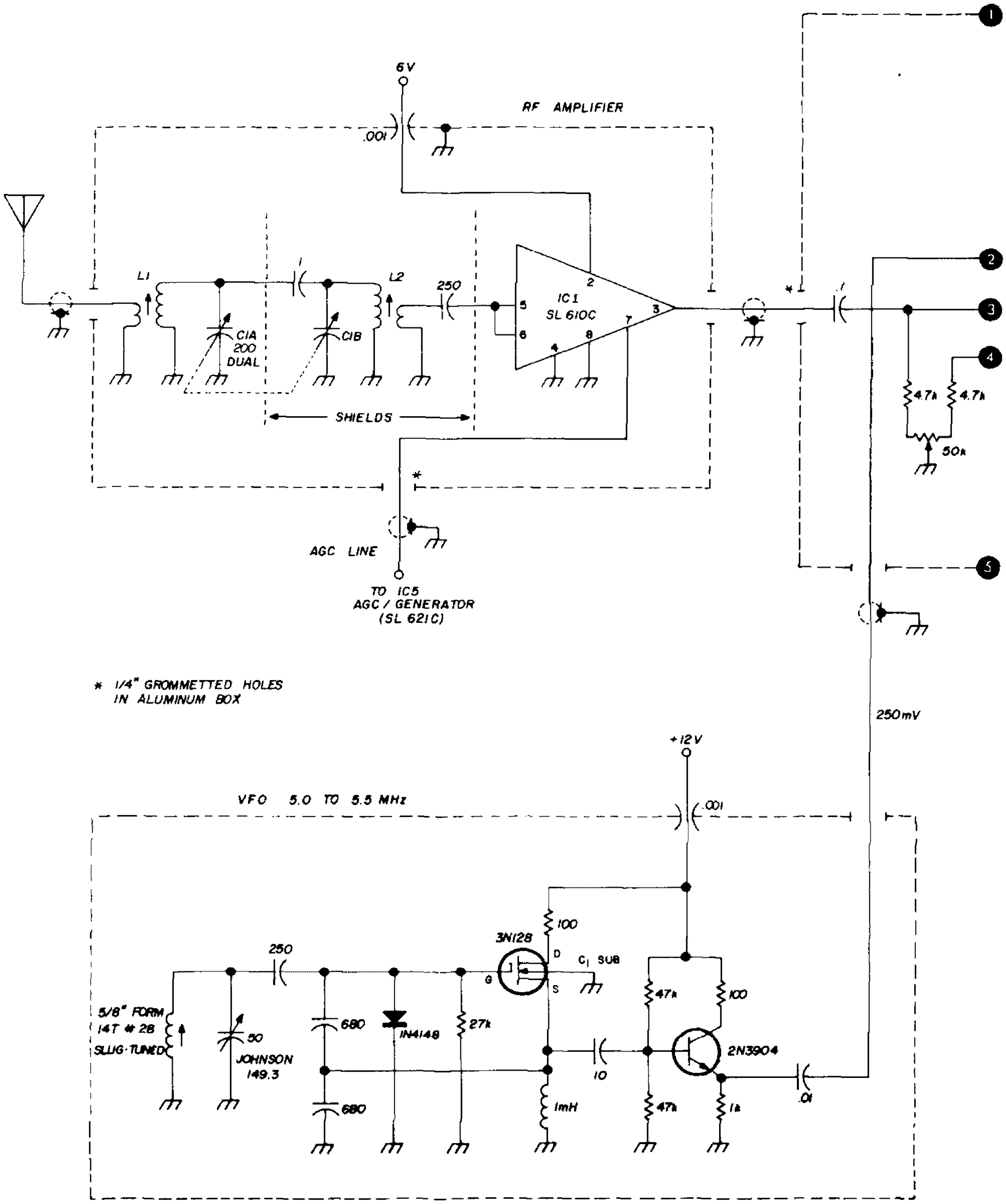
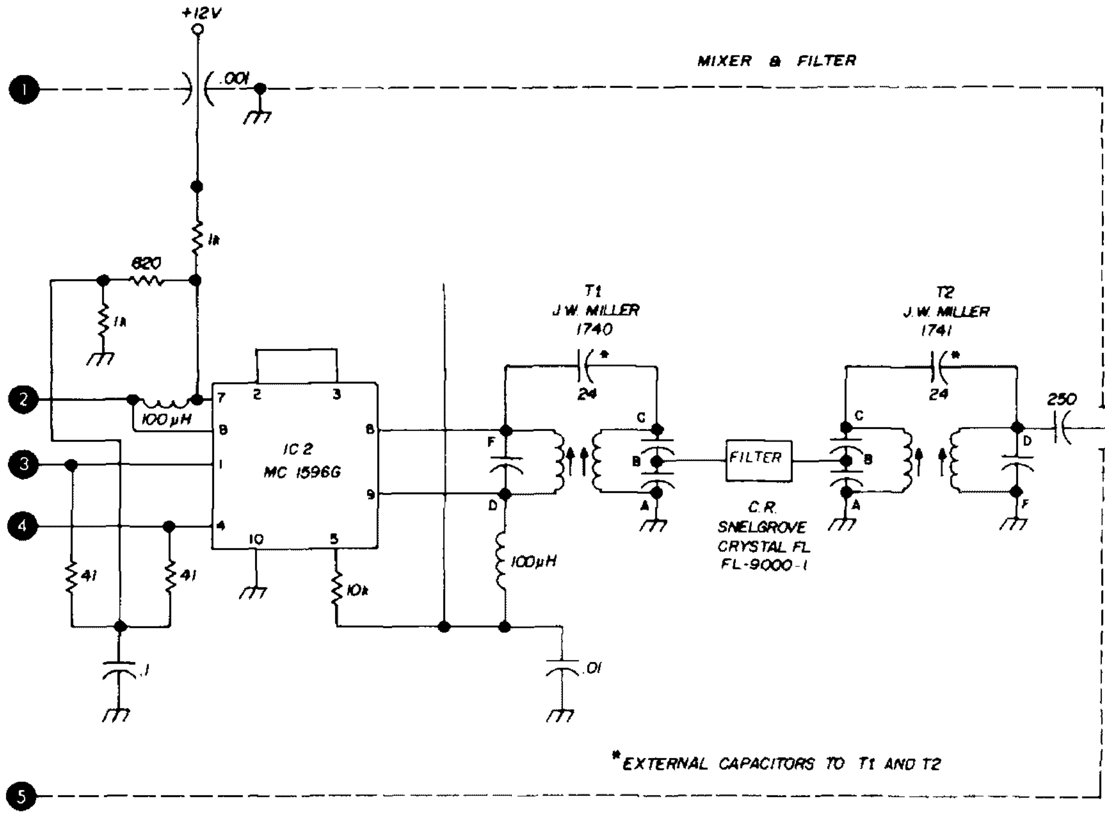
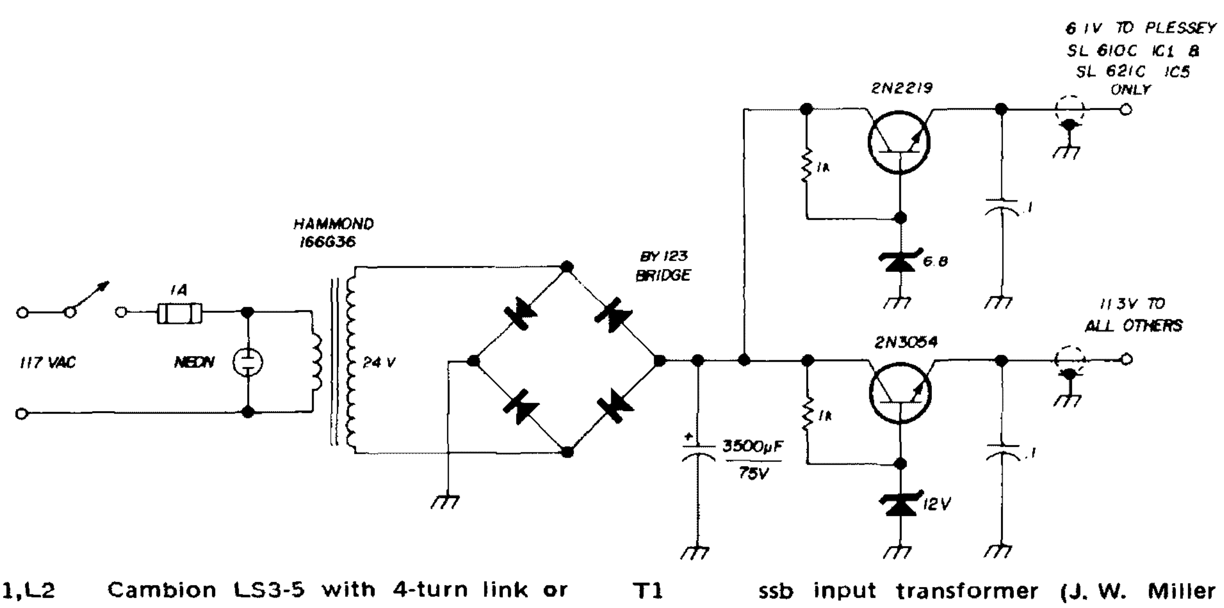


fig. 3. Circuit diagram of the final design integrated-circuit 80-meter receiver. Power-supply circuit is shown on page 9. Bfo crystals Y1 and Y2 are furnished with the crystal filter. Author used Snelgrove FL-9000-1 crystal filter. However, KVG XF-9B or McCoy Silver Sentinel will work as well.

and output transformers, the balanced mixer configuration is the same as that described by K7WQR. The Miller transformers have low loss, provide sharp tuning and appear to match the crystal filter very well. The 24-pF capacitors across the transformer windings provide increased gain. These transformers eliminate the complex problem of matching the input and output impedance of the crystal filter. Incorrect impedance matching severely distorts the passband of the filter; homebrewing these transformers is much too difficult for the average experimenter.

The 90-dB dynamic range of the

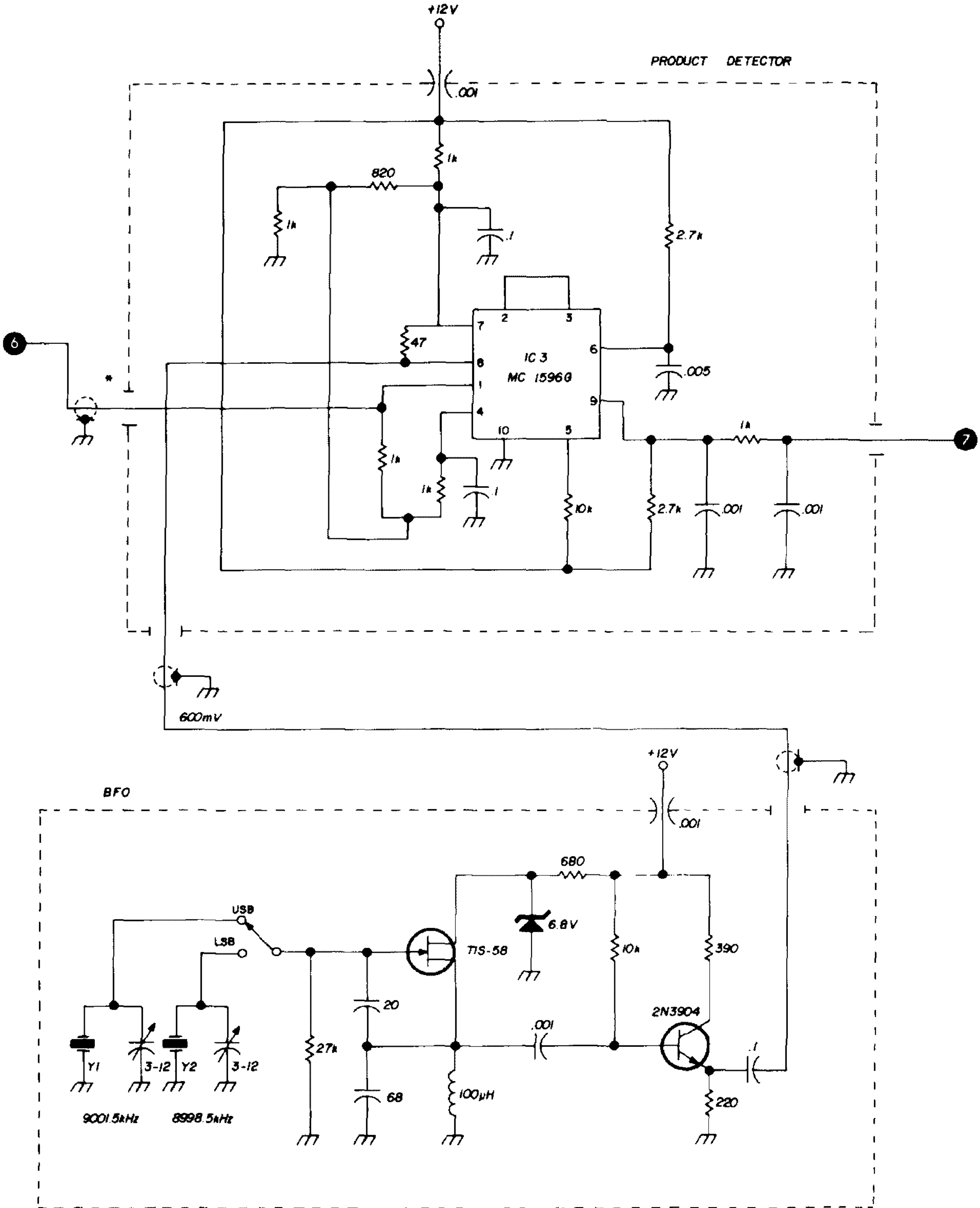




L1,L2	Cambion LS3-5 with 4-turn link or 25 turns no. 22 enameled on Amidon T50-6 toroid form with 3-turn no. 22 link	T1	ssb input transformer (J. W. Miller 1740)
L3	14 turns no. 28, 5/8" diameter, slug tuned.	T2	ssb output transformer (J. W. Miller 1741)

MC1596G integrated circuit makes the device relatively insensitive to input levels. This fact, coupled with the 3-μV sensitivity and high conversion gain, makes the MC1596G the logical choice for the balanced mixer stage.

The only adjustments in the mixer/filter module are T1, T2 and the 50k null-adjustment pot. Transformers T1 and T2 are tuned for maximum signal. The null-adjustment pot is set for maximum 9-MHz signal at pin D of T2. These

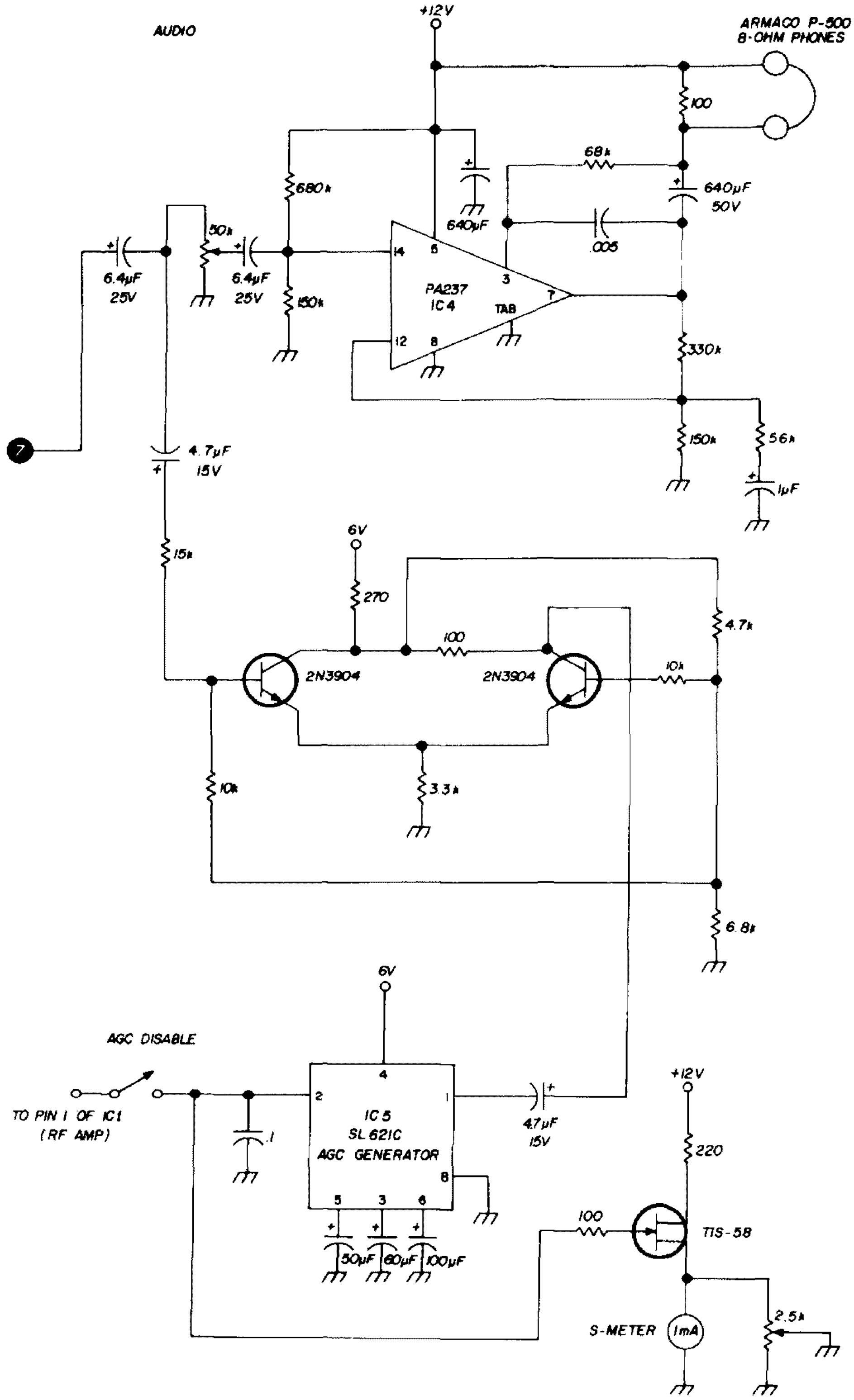


adjustments are quite simple as they are rather well defined.

vfo

The vfo is basically the same as the circuit presented in the 1969 edition of The Radio Amateur's Handbook.[3] This vfo is extremely stable and the output waveform is clean and symmetrical. A bandpass filter is not required on the output because the high common-mode rejection of the IC balanced mixer (typically 85 dB) eliminates any unwanted sidebands present in the vfo signal.

With this circuit vfo drift over a period of several hours was on the order of a few cycles per second. This far exceeds the requirements for this receiver. I have used this circuit several times, and the purchase cost of the 3N128 mosfet is well worth its performance and stability. How-



ever, to obtain good stability it is important to use a good quality double-bearing air-variable capacitor along with rigid mechanical design.

product detector

The product detector circuit is essentially the same as that described by K7WQR. High-level signal performance and sensitivity is excellent; conversion gain is high, typically 24 dB. In fact, performance of this circuit is so good, VE3GFN has discarded his hot-carrier diode product detector[4] and replaced it with this one.

audio

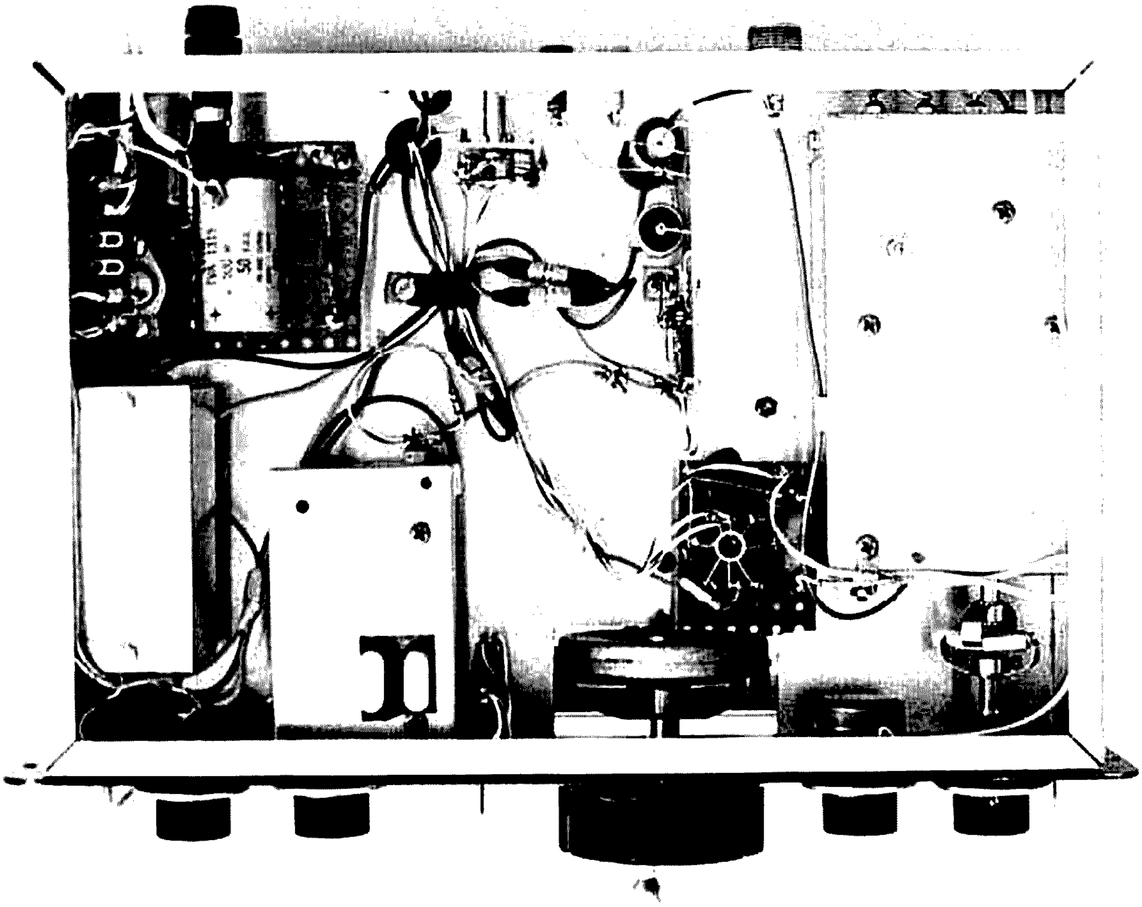
The audio system uses another IC, the General Electric PA237. Although this IC should be used with a power supply of 24 volts for its rated output power of 2 watts, with a 12-volt power supply it

drives adequately a pair of Armaco P-500 stereo headphones. A less expensive audio system would probably result from using a consumer-rated audio-power IC such as the Motorola MFC9010.

agc

The agc system is built around the SL621C agc generator manufactured by Plessey Microelectronics.* This IC is designed to be used with the SL610C rf amplifier in ssb applications. As with other advanced agc systems, this integrated circuit generates the agc voltage directly from the detected audio waveform, provides a "hold" period to maintain the agc level during speech pauses and is immune to noise interference. When the SL621 agc generator is used in conjunction with the SL610 rf amplifier, audio level is maintained within 4 dB for a 110-dB range of receiver input signal. In my receiver the audio level holds within 4 dB for rf input signals up to 150 mV rms.

If you consider using the SL621C I recommend that you read the excellent article by W3TNO.[5] The input to the SL621C must be held to approximately 10 mV rms. In the circuit in fig. 3 I have used W3TNO's method of accomplishing



Under the chassis. From left to right, front, product detector, bfo, agc generator and rf amplifier. Power supply is upper left. The audio output board is on edge in the upper center.

*Plessey integrated circuits are available from Plessey Electronics Corporation, 170 Finn Court, Farmingdale, Long Island, New York 11735. The SL610C rf amplifier and SL612C i-f amplifier are $5.65 each; the SL621C agc generator is $8.30.

The following semiconductors are available from Circuit Specialists Company, Box 3047, Scottsdale, Arizona 87257. MC1496 ($3.90), TIS58/HEP802 ($1.59), 2N3904/HEP736 ($1.28), 2N2219 ($1.38), 2N3054/HEP703 ($2.75), PA237 ($4.30), 3N128 ($1.74), 12-volt Z0222 zener (.70), 6-volt Z0215 zener (.70) and 1N4148 (.40). Please add 25c for shipping.

this: An emitter-coupled clipper to prevent the input capacitor from charging, and a resistive voltage divider to keep the audio signal at the 10 mV level required by the IC.

power supply

The two voltage requirements of 6 and

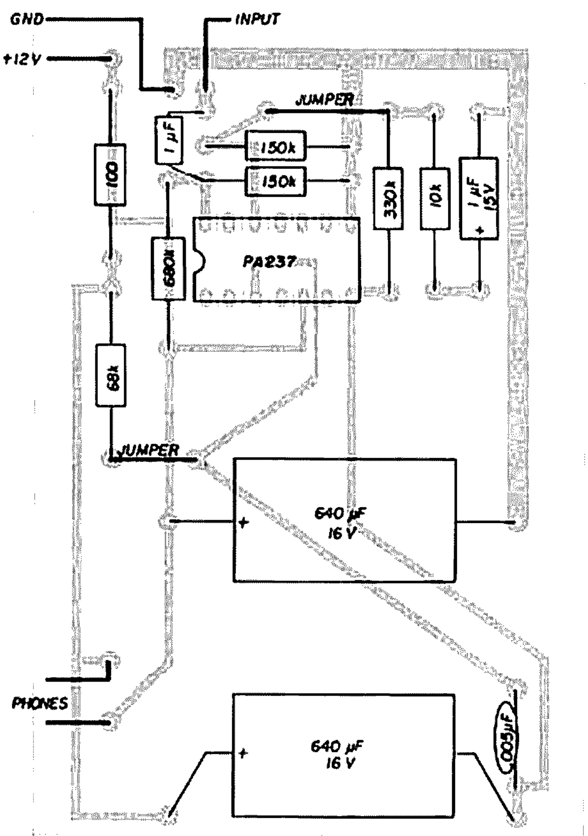


fig. 4. Printed-circuit layout for the 1-watt integrated-circuit audio power amplifier.

12 volts are derived from series-type voltage regulation (fig. 3). This method is quite good if a large amount of filtering is included in the circuit. However, for better regulation and less ripple you should consider one of the integrated-circuit voltage regulators that are available.

performance

The sensitivity of the complete communications receiver is 0.5 μV for 10-dB signal-plus-noise-to-noise ratio; this was measured with a Boonton type 65B signal generator. No cross modulation was detected for input levels up to 100 mV rms. Frequency readout (with the Eddystone 898* dial) is to 1 kHz; frequency error over the 500-kHz tuning range is 18 kHz.

***The spinner-type main-tuning knob shown in the photographs is not standard; it was machined by VE3ELP.**

conclusions

I plan to use other versions of this basic design in future homebrew receivers. One possible approach is to feed the antenna directly into the first mixer and use agc in the i-f section as in the W7ZOI design.[7] The Plessey SL612 integrated-circuit i-f amplifier is compatible with the SL621C agc generator and looks like an ideal choice. I am presently working on an all-band converter using the MC1596G balanced-mixer IC that will use this receiver as a tunable i-f from 3.5 to 4.0 MHz.

I would like to thank M. Goldstein, VE3GFN, and A. Martin for the photographs; I would also like to thank J. Riach, VE3DSR, for his constructive criticism.

references

1. R. Hejhall, K7QWR, "An Integrated-Circuit Balanced Modulator," *ham radio,* September, 1970, page 6.
2. W. Sabin, WØIYH, "The Solid-State Receiver," *QST,* July, 1970, page 35.
3. "A Stable FET VFO," The Radio Amateur's Handbook, A. R. R. L., 1969 edition, page 188.
4. M. Goldstein, VE3GFN, "Hot-Carrier Diode Product Detector," *ham radio,* October, 1969, page 12.
5. J. Gotwals, W3TNO, "Solid-State Receiver for 160 Meters," *ham radio,* December, 1969, page 6.
6. Plessey Microelectronic SL600 specification sheets.
7. W. Hayward, W7ZOI, "A Second Generation Mosfet Receiver," *QST,* December, 1970, page 11.
8. G. Jones, W1PLJ, "A Versatile Solid-State Receiver," *ham radio,* July, 1970, page 10.
9. J. Fisk, W1DTY, "Homebrew Single-Sideband Transceiver," *ham radio,* June, 1969, page, 8.
10. H. Olson, W6GXN, "Solid-State Power Supplies," *ham radio,* February, 1970, page 25.
11. R. Fischer, DL6WD, "An Engineer's Ham-Band Receiver," *QST,* March, 1970, page 11.
12. W. Scherer, W2AEF, "Receiver Signal-Handling Capabilities," *CQ,* January, 1970, page 39.

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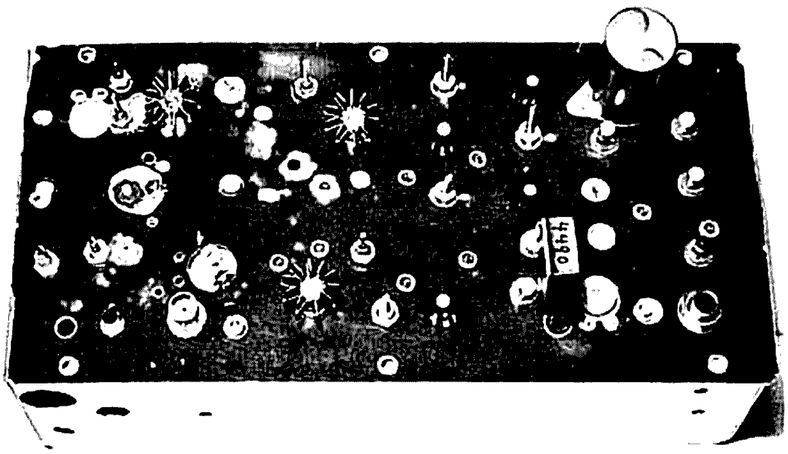


solid-state two-meter fm transmitter

This high-performance phase-modulated 144-MHz transmitter features 8-watts output with a 12-volt power supply

Frank Jones, W6AJF, 850 Donner Avenue, Sonoma, California 95476

This two-meter **fm** transmitter is suitable for operation through the fm repeater stations which are located on a great many mountain tops. With an auto-battery power supply of 12 to 14 volts carrier output measures from 2 to 3 watts. A 24-volt supply to the final 2N3632 power stage will provide 7 to 8 watts of fm output. The 2N3632 transistors are not the most desirable type, but surplus devices were less expensive than the newer types which were available a year or so ago. Some 2N3632 transistors run wild and burn out easily when used with a power supply greater than 15 volts.

Phase modulation, converted to fm, is much easier to handle than a-m with transistor systems, and carrier output is about twice as high as that for a-m. The collector supply with amplitude modulation has to be reduced greatly with most power transistors to operate within the maximum allowable peak voltages. Ssb has the same problems. These modulation systems require power transistors with good linearity and peak voltage ratings more than four times the value of dc

supply.[1] Phase or frequency modulation requires that the peak voltage under load conditions be less than twice the dc supply value. This permits supply voltages two times higher and results in 2 to 4 times as much rf carrier output.

In an overall system with a good fm receiver (not slope detection) fm can compete with ssb in effectiveness; it may even surpass ssb for difficult paths on 144 MHz. Also, ignition and some power noises are suppressed much better by fm receivers since the limiters in an fm receiver which remove a-m also knock off noise pulses at the carrier level. This cannot be done in an a-m or ssb receiver without some pretty sophisticated i-f noise blankers.

This transmitter has been in intermittent service for over a year with a repeater which is behind a small mountain range. It is not line of sight, and the "knife-edge" diffraction effects are very poor in the direction of my station. The previous 1-watt fm transmitter wouldn't do the job even with a 10-dB gain antenna. This transmitter usually does the job.

phase modulator

The phase modulator uses a pair of varactor diodes — rectifiers normally used for small power supplies. The 1N645 rectifier has some variable capacitance effect with variations of audio voltage when the diodes are reverse biased with a few volts. Nearly any small power rectifier with very high back resistance will work in this circuit. Some types have much higher resting capacitance and require a smaller shunt inductance to resonate at 4.6 MHz.

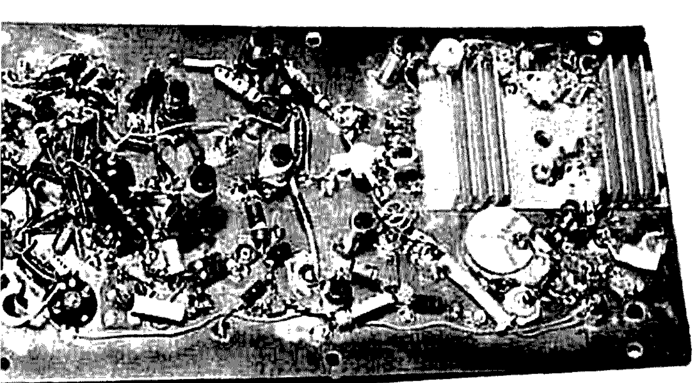
The 1N645s in series work with a slug-tuned coil (from an older type broadcast receiver) since the total shunt capacitance is only a few picofarads. The untuned crystal oscillator is lightly loaded through a 2-pF capacitor. The crystal can be zeroed in on a desired frequency if it is slightly off by means of a variable capacitor from one side of the crystal to ground.

The phase modulator LC circuit should be used to drive a high input impedance class-A amplifier through a 1- or 2-pF capacitor. Actually, it can be used to drive a frequency doubler without too much distortion if the rf output requirements are moderate. In the circuit shown in fig. 1 frequency doublers are used throughout; in the first few doublers the power requirements can be met with small surplus transistors of the 2N706 or 2N708 class. The 2N5188 power transistor can be made to work up to 148 MHz but it will not put out much over 250 or 500 milliwatts since a 12-volt power supply is near its maximum rating. The 2N5188 is in the half-dollar price range from RCA distributors.

This type of frequency-doubler drive circuit has proven to be much superior to inductive-coupled circuits. The capacitance dividers step the tuned-circuit impedance down to 50 ohms or so to work into the base of the next stage. The large capacitance across the base acts as a low impedance of a few ohms at the doubled output frequency. Without a low-impedance path from base to emitter the collector-to-base capacitance will feed back enough energy to greatly reduce output power.

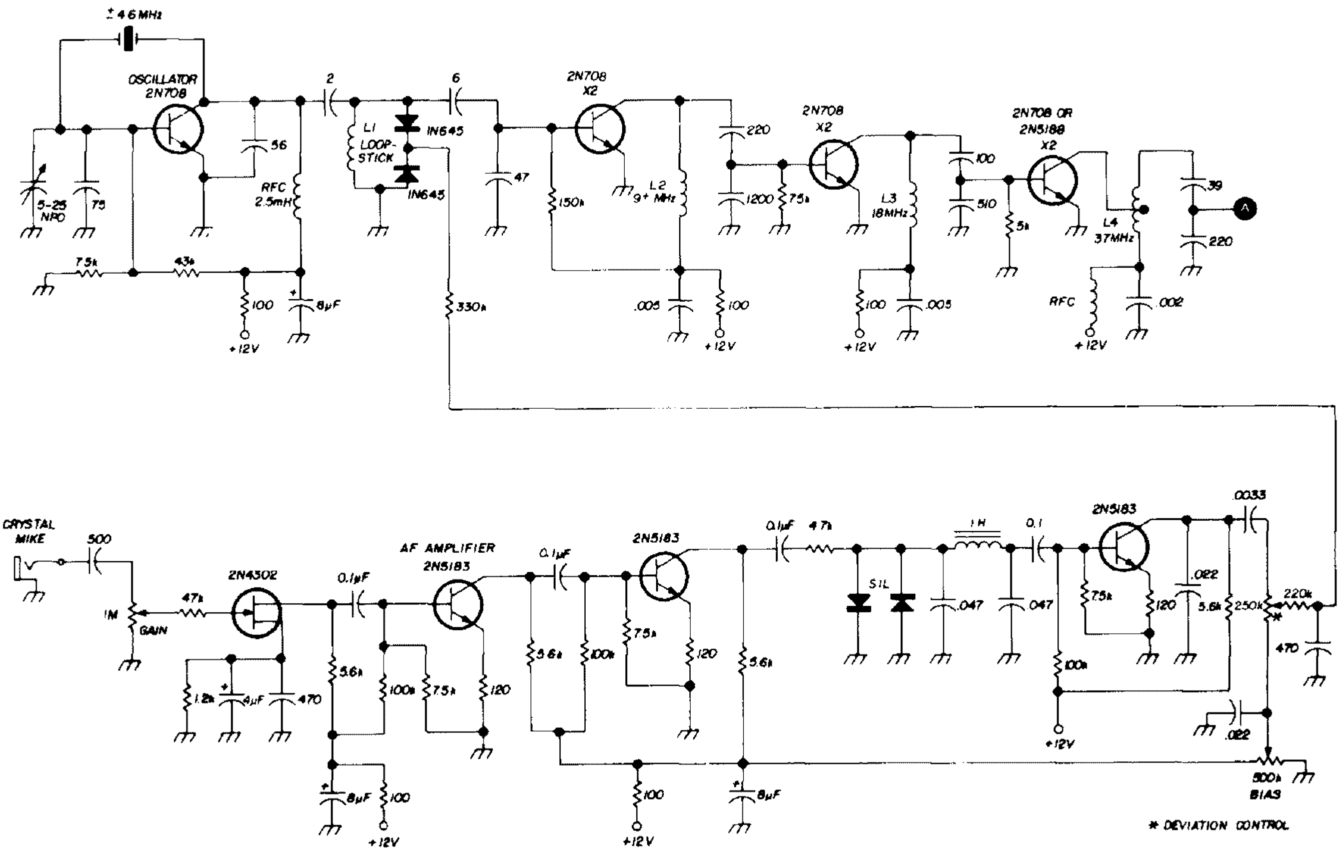
This frequency multiplier design saves dc power and transistor heating, and

Inside the fm transmitter. Rf power output transistor is stud-mounted to the large heat sink on the right. Mounting chassis is 3x5x10 inches.



permits considerable latitude in choosing transistors. For maximum efficiency the transistor should have a cutoff frequency rating of 5 to 10 times the operating frequency. The 2N5188 is operating too close to its cutoff frequency for good efficiency as a two-meter doubler, but some devices work reasonably well as two-meter amplifiers.

The lower-frequency doubler stages are not as critical, so I moved the transistors around for best overall drive to the final amplifier. One 2N5188 worked fairly well in the 144-MHz 2N3553 stage (at 1/10 the cost) with only a 12-volt power supply; the 2N5188 will not work at higher supply voltages. However, this transmitter is normally used with a 12-volt storage battery so the 2N5188 is used most of the time in the 144-MHz stage to save the more expensive 2N3553 for other uses. (The 2N3553 makes a good final amplifier with a 24- or 28-volt supply providing 5 to 7 watts output.) With a 12-volt supply power output is less than 2 watts, somewhat inferior to the 2N3632 which costs less on the surplus market.



rf amplifiers

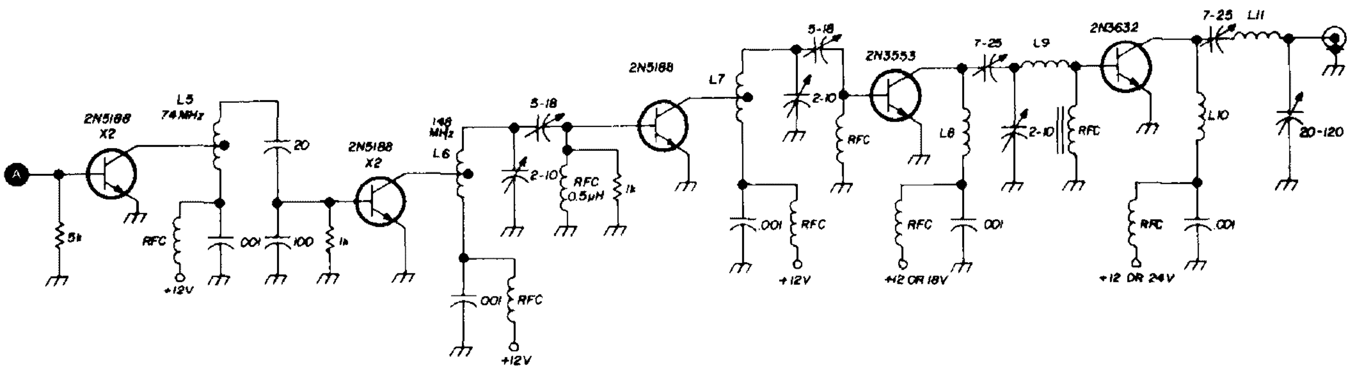
The rf amplifier circuits were chosen from a standpoint of less parasitic oscillation and transistor burnout than other circuits I've tried in previous transmitters. The little power transistors are real monsters unless the base-to-emitter path is just right. A small rf choke consisting of 5 or 6 turns of wire on a lossy 1/8-inch diameter ferrite rod makes about the best dc path from base to emitter. If the rf driving power is large enough, a 10-ohm carbon-composition or carbon-film resistor can be used for this dc path in place of the ferrite-core rf choke.

The more rf and dc power into and out of a transistor, the lower the input and output impedances. You can expect values of 2 to 10 ohms in the output stage. This makes it difficult to match impedances to the driver and the antenna circuits.

As shown, with only one tuned circuit,

harmonic output is high. However, I use a dual-cavity antenna filter on all transmitters and receivers between the antenna coax lead and the coaxial relay. This reduces the harmonic content of all

ing is needed in this stage. The 2N5183 and 2N3565 transistors used in the other audio stages are low priced high gain bypolar types. A diode speech clipper was added to keep the average modulation



L1	100-250 μH loopstick
L2,L3	20 turns no. 26 enameled on ¼-inch form, 3/8-inch long
L4	10 turns no. 22 enameled on ¼-inch form, 3/8-inch long, center-tapped
L5	7 turns no. 20 enameled, center-tapped, ¼-inch diameter with brass slug
L6,L7	5 turns no. 20, tapped 2 turns from ground, on ¼-inch form, 3/8-inch long
L8	4 turns no. 20, 3/16-inch diameter, ½-inch long
L9	6 turns no. 20, 3/16-inch diameter, ¾-inch long
L10	3½ turns no. 20, 3/16-inch diameter, ½-inch long
L11	3 turns 1/8-inch copper ribbon, 3/8-inch diameter, ½-inch long

fig. 1. Schematic for the 8-watt two-meter fm transmitter. Small decoupling rf chokes in the 12-volt supply lines are approximately ½-μH. The ferrite-core rf chokes in the last two stages consist of 6 turns no. 20 enameled on a 1/8-inch diameter powdered-iron rod from a 455-kHz i-f transformer.

144-MHz transmitters and reduces image response when receiving. By using an unloaded Q of 1500 to 2500 in this antenna filter, loaded down to a Q of 15 or 20 by the 50-ohm terminations, signal loss is quite small. Two tuned circuits provide about 40-dB harmonic suppression with less loss than with one higher-Q circuit loaded less heavily.

audio

The audio system in this transmitter is designed for high-impedance crystal or ceramic microphones. The fet pre-amplifier (2N4302 or MPF103, etc.) has a high input impedance. Rf feedback in a two-meter transmitter is always a problem so some rf filtering and bypass-

level as high as possible. The frequency components above 2000 Hz are reduced by a low-pass filter consisting of a small 1-H choke and a pair of 0.047-μF capacitors. The higher audio frequencies must be attenuated in order to convert from phase to frequency modulation.[2] The phase modulator diodes require only a volt or two of audio in the lower voice frequency range, and a small fraction of a volt at 2500 Hz. The RC values in the last audio stage aid this affect.

references

1. Paul Franson, WA7KRE, "How to Use RF Power Transistors," *ham radio*, January, 1970, page 8.
2. Les Cobb, W6TEE, "Modulation Standards for VHF FM," *ham radio*, June, 1970, page 16.

ham radio

crt intensifier for RTTY

A simple circuit for intensifying the RTTY monitor scope during receive

Heard Lowry, K4VFA, 915 Madison Street, Manchester, Tennessee 37355

Radioteletype is a very interesting facet of amateur radio, and equipment is often designed for the highest degree of automation. When using an oscilloscope as a tuning aid there is no problem of beam brilliance as long as the beam is being deflected by the incoming mark and space signals. The beam can be adjusted, using the intensity control, for proper brilliance with signal. However, a problem arises during RTTY operation when no-signal periods occur—the beam is not moving, and screen brightness becomes very intense. Unfortunately, the high intensity spot will burn the screen if the intensity control is not turned back.

The inconvenience of monitoring beam brilliance and continually making adjustments to save the cathode ray screen led me to look for a system that would accomplish this function automatically.

The desired circuit for controlling cathode-ray-tube bias should OR the mark and space signals so that if either is present the grid bias on the cathode ray tube is lowered, and the signal pattern is displayed.

Two requirements are necessary for controlling cathode-ray-tube bias. First, the bias must not get low enough to cause the electron beam to bloom (get out of focus). The beam focus point changes when the beam current passes a certain point, and this must be avoided. The maximum positive voltage that the control circuit can apply to the grid of the tube must be limited to a maximum that will not cause blooming of the beam.

Second, the control circuit should have some degree of proportional control when the incoming mark and space signals begin to fall below the level at which limiting takes place in the RTTY converter and the display on the cathode ray tube screen begins to shrink. As the display begins to shrink and beam trave speed is reduced, screen brilliance wil increase due to the reduced area o illumination. The increase of brilliance with reduced display size points out the desirability of a brightness control that is proportional to display size when signa levels fall below the limiting levels of the RTTY converter.

the circuit

A circuit which will provide brightnes control of the cathode ray tube durin typical RTTY signal conditions is show in fig. 1. The intensifier circuit is con

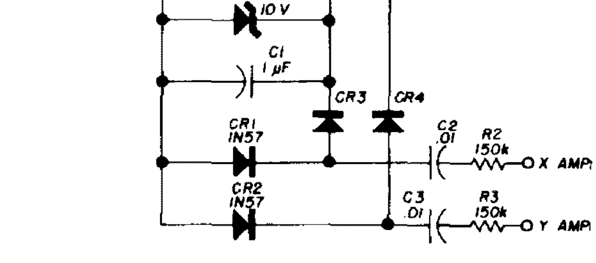


fig. 1. Schematic of the CRT intensifier. Value of CR5 is discussed in text.

nected between the brightness control and the cathode-ray-tube control grid. It is activated by space and mark signals from the oscilloscope's X and Y amplifiers. Loading on the X and Y amplifiers is reduced by the 150k resistors R2 and R3. Coupling to the X and Y channels is provided by C2 and R2 to the X amplifier and C3 and R3 to the Y amplifier.

operation

The dc intensification voltage is developed across R and C. The peak intensification voltage is set by the 10-volt zener diode, CR5, and limits this voltage to a safe value which will not cause blooming

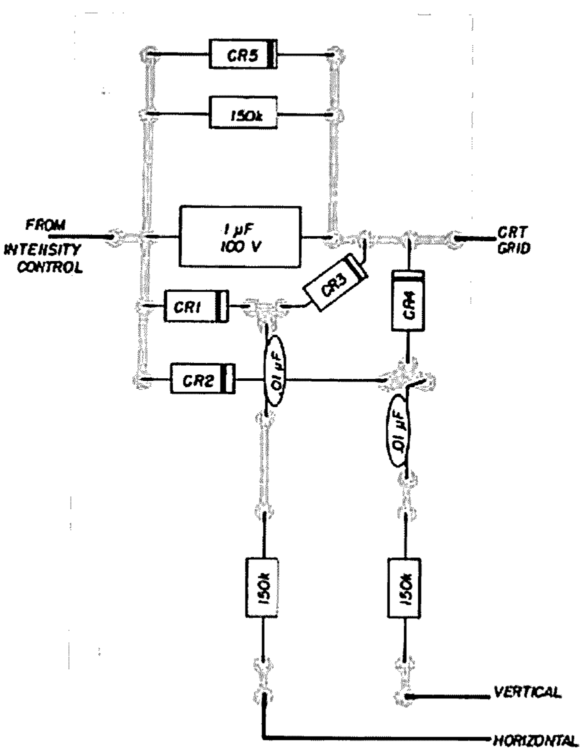


fig. 2. Full-size printed-circuit layout for the CRT intensifier.

of the beam during intensification. Selection of the zener diode as well as suggestions for particular applications will be discussed later.

The rectifier circuits from the X and Y oscilloscope channels consist of voltage-doubler circuits. The X circuit is made up by CR1, CR3, and C2, with C1 and R1 common to both the X and Y channels. The Y circuit consists of CR2, CR4 and C3 with C1 and R1. The X and Y signals feed into their respective doubler circuits; the highest signal, mark or space, sets the voltage level until the zener diode begins to conduct and limit the voltage to a peak brightness level. The time constant for the brightness level is determined by C1 and R1. By varying C, the length of time is controlled during which intensification is extended.

construction

To determine the required zener voltage for particular scope the following procedure is recommended. A high-impedance voltmeter is connected between the grid number 1 and the cathode of the CRT. (*Caution.* High voltages are usually involved, so do not change connections or touch the voltmeter while power is applied.)

The first measurement is obtained by applying power to the scope without mark or space input signals. (This should produce a spot.) Reduce intensity until the spot is completely extinguished; then read and record the voltage. The next reading is made with both mark and space signals applied, the CRT showing a cross display. Adjust the brilliance until a pleasing display intensity is obtained; then read the voltmeter and record this reading. To obtain the zener voltage rating subtract the second reading from the first; a quarter-watt zener diode with this voltage rating will provide good results.

The CRT intensifier can be built on a small printed-circuit board or Vector-type perforated board. Or, you may mount the components on two or three terminal strips which have been installed at convenient spots on the scope chassis. Fig. 2 shows a printed-circuit layout which includes mounting holes. This layout is small enough so that no trouble should be experienced in finding a suitable location within your scope. The board can be vertically mounted with L-shaped brackets. If the circuit board is mounted inside the scope give proper consideration to lead length and proximity to other components.

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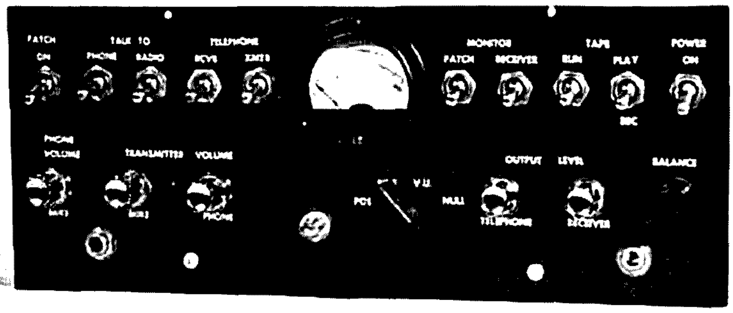

superior phone patch

A legal phone patch with many desirable features including complete operator control

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Hybrid circuits to interconnect the amateur station and a commercial telephone circuit are not uncommon, both in kit form and assembled. Most such phone patches, however, have a number of disadvantages. These include the necessity of using the telephone handset, which frequently over-drives the transmitter; complex prodecures to go from normal operation to patch and return; and unsatisfactory vox operation.

features

The unit described in this article overcomes the problems mentioned above and has additional desirable features. The operator can talk to either party – telephone or radio – together or selectively, using the station microphone and headset or speaker. He is in complete control, with ability to cut off or override either party at any time. Changeover from normal operation to patch and back is simple and fast. Finally, it is possible to record all three parties' conversations and play the recording back to telephone, radio, or local monitor.

A block diagram of the system is shown in **fig. 1.** Even without the tape recorder connections, which are omitted, it is apparent that each input – transmitter, telephone, and monitor – receives signals from two sources; and some form

of hybrid or other isolation is required to keep the signal originating at one source from feeding back to the other. However, the only place where input and output circuits have common terminals is the phone, so this is the only place a hybrid is required. The rest of the isolation can be accomplished with amplifiers, indicated by triangles in the diagram.

The original version of this patch used transistors and other discrete components. The present model employs IC operational amplifiers, resulting in a simpler circuit at no greater cost. The amplifiers are Motorola MC 1439Gs – internally protected against input overload, latch-up, and output short-circuit; and they're relatively inexpensive.

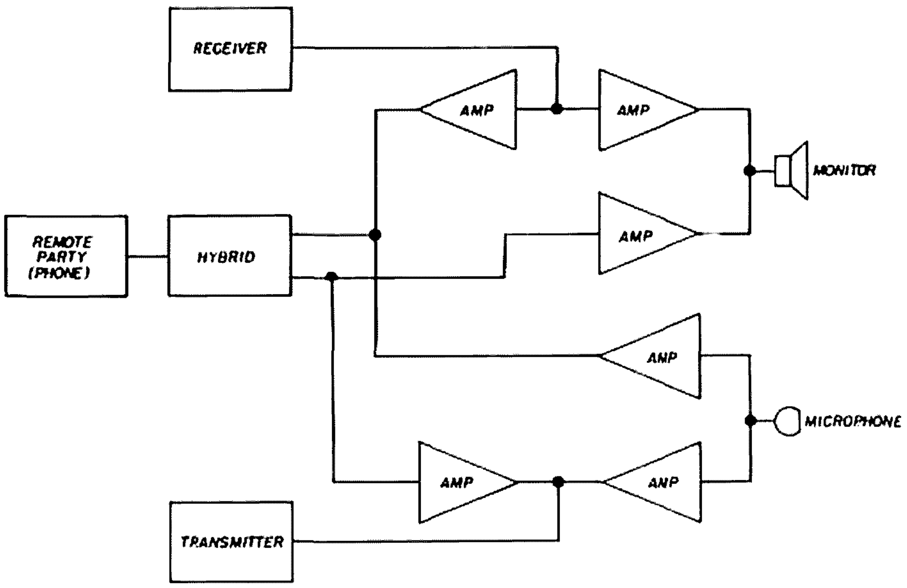
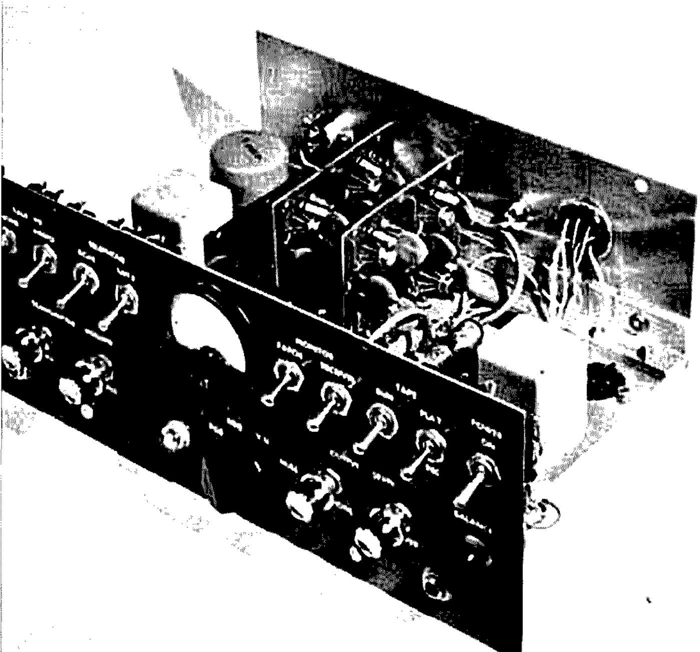


fig. 1. System block diagram. Each input receives signals from two sources; feedback is eliminated by hybrid circuit and amplifiers.



Connections for a typical audio amplifier are shown in **fig. 2.** The model shown is an inverting amplifier, with signal input at the inverting (-) lead and the noninverting (+) lead grounded. It is stabilized by 0.1 μF bypass capacitors at leads 4 and 7 and a compensation network between leads 1 and 8. Values shown are recommended by the manufacturer for unity gain; they work well for all gains and frequencies encountered in the phone patch. This compensation is applied to all seven amplifiers in the patch but is omitted from the schematic (fig. 3) for simplicity.

Gain is determined by the ratio of R_F to R_I; and a feedback capacitor, C_F, is added to three amplifiers to suppress unwanted high-frequency response.

circuit

In **fig.** 3, T1 and T2 form the hybrid. I use a pair of UTC transistor/line transformers, type A-22, which have split secondaries with 250 ohms impedance on each half, and 500-ohm center-tapped primaries. The secondaries are cross-connected so that a signal injected into one

primary is balanced out in the other transformer. These transformers work very well, but those wishing to build a similar unit from scratch may want to try a transformer with a higher-impedance primary for T1, and possibly eliminate the two 470-ohm resistors I had to put in series with the amplifier outputs to make them work properly.

T3 is a one-to-one isolation transformer with from 500- to 1000-ohm impedance each side and a third winding that provides a signal for the tape recorder input and VU meter.

The pad between the hybrid and T3 presents a relatively constant impedance to the hybrid, facilitating balancing with

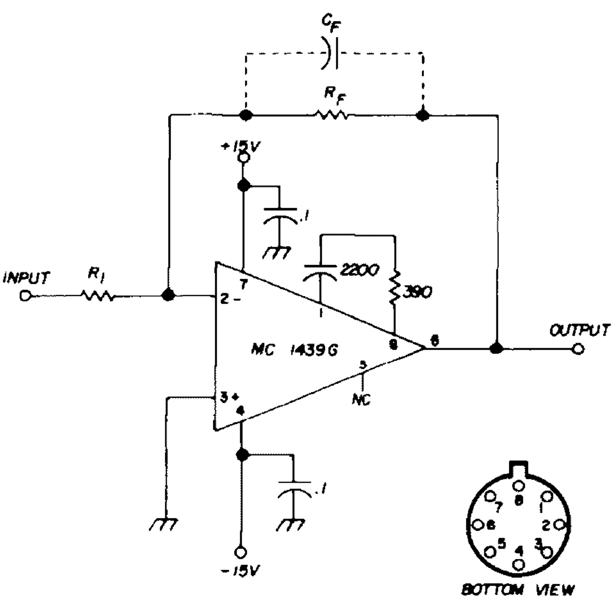


fig. 2. Terminal connections and compensation for a typical IC audio amplifier.

the 1k variable resistor. Good balance is essential if vox operation is desired.

amplifier considerations

As mentioned earlier, the power supply and compensation connections to the amplifiers have been omitted from the schematic for clarity. Amplifiers 1, 4, and 6 are simple inverting models, with adjustable gain controlled by varying their feedback resistances. The .001-μF feedback capacitors on amplifiers 1 and 6 suppress an approximately 25-kHz oscillation that appeared when the unit was connected to one of my station receivers; others may not find them necessary.

Amplifier 5 is similar; but instead of adjustable gain, it has an RC network in its feedback circuit to make it an active filter as well as an amplifier. Signal level from the receiver to the phone line is controlled with the gain controls on the receiver.

Tariffs of many Bell System telephone companies, in addition to limiting the strength of the signal that may be transmitted over their lines, require that devices connected to the lines through their voice coupler have minimal output at 2600 Hz and above.[1, 3] The circuit shown has a maximum voltage gain of 10 at approximately 1000 Hz and practically no gain at 400 and 2000 Hz. This narrow bandpass admittedly sacrifices some voice quality, but intelligibility seems to be generally good. The circuit eliminates a lot of heterodynes and other interference and keeps the telephone company happy as well. Those desiring a broader bandpass, while still meeting phone company requirements, might try a two- or three-stage biquadratic filter[2] in place of amplifier 5.

Amplifiers 2 and 3 were originally constructed similarly to the others, but two problems resulted. First, with the microphone and the feedback resistor both connected to the inverting input of amplifier 2, the receiver output to T1 from amplifier 5 fed back around amplifier 2 to the input of amplifier 3 and the transmitter, which defeated the purpose of the hybrid. To correct this, amplifier 2 was connected as a noninverting amplifier, with the microphone input to lead 3 and feedback to lead 2.

Furthermore, the high gain needed in amplifier 2 to amplify the microphone signal to phone-line level makes the shielding of the microphone leads to amplifiers 2 and 3 critical. The circuitry used permits mounting both the input and feedback resistors directly on the circuit boards, minimizing lead length and simplifying shielding.

Amplifier 7 is a meter amplifier. Positions 1 and 2 of the meter switch (S11)

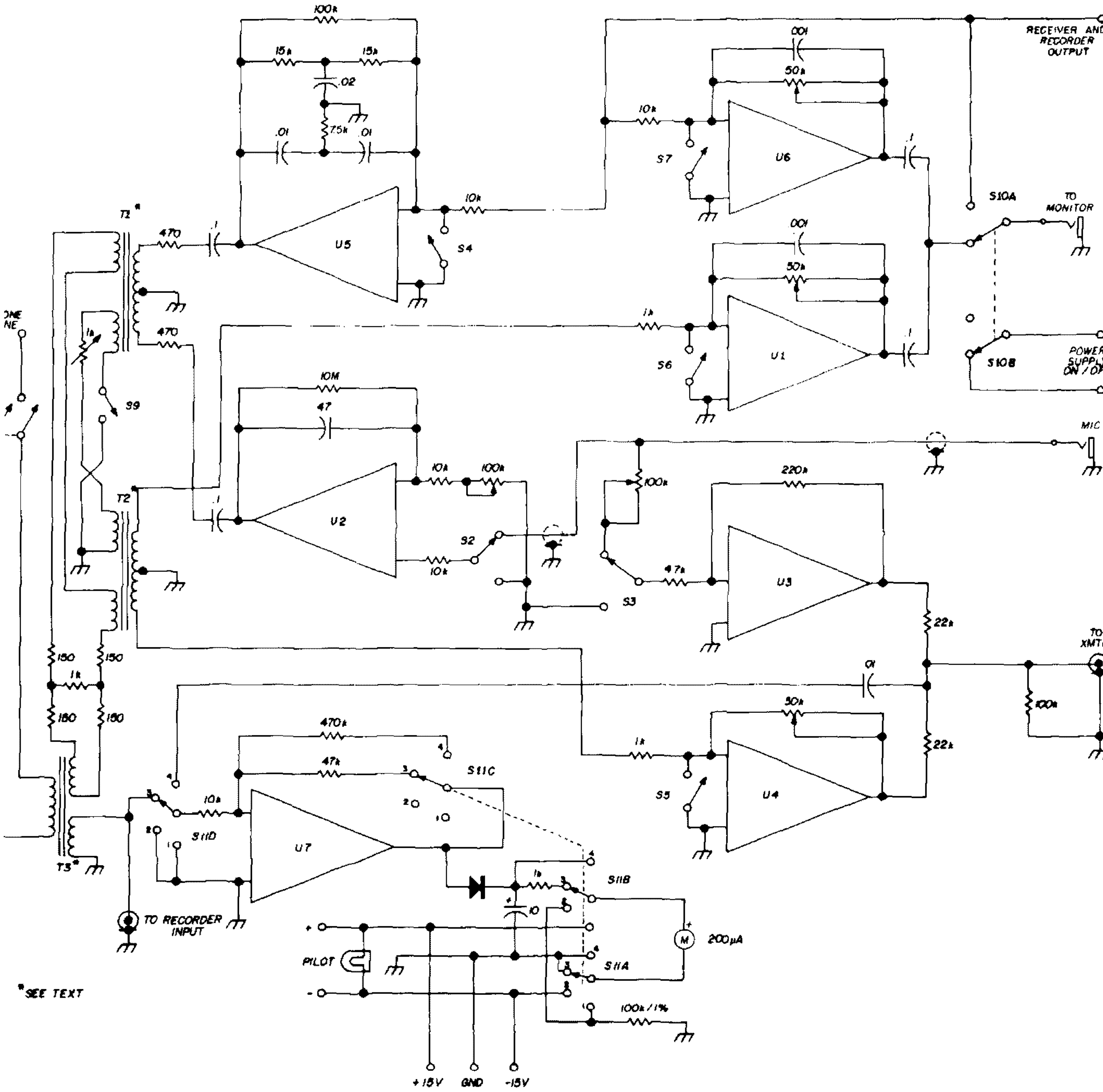


fig. 3. Phone patch schematic. Transformers T_1, T_2 form the hybrid between telephone and patch, which have common input and output circuits.

read power-supply voltage; in position 3 the amplifier samples the signal on the phone line, making it possible to monitor the level of the outgoing signal. In position 4, used to balance the patch circuit, the meter shows the signal level at the transmitter input.

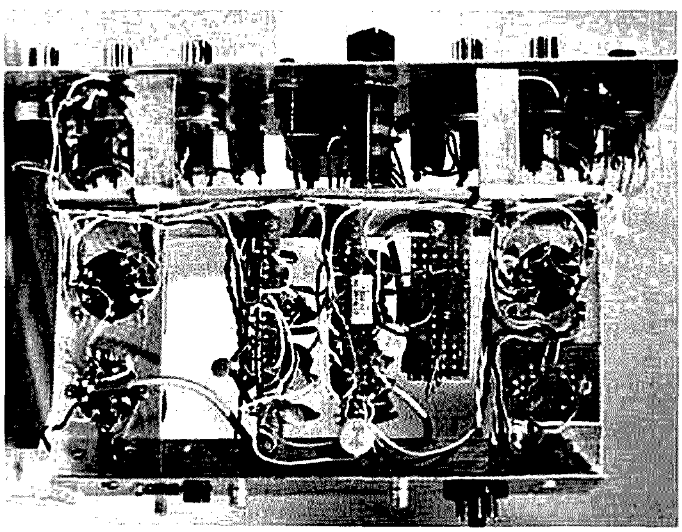
controls

In addition to the six potentiometers, there are ten toggle switches and the rotary meter switch on the control panel. The PATCH ON/OFF switch, S1, connects the patch to the telephone circuit. Switches S2 through S7 are ON/OFF switches for the six audio amplifiers and determine which patch functions are active. S8, not shown in the circuit diagram, is a tape recorder START/STOP switch, like the pushbutton on some tape recorder microphones. S9 unbalances the hybrid by opening one side to permit tape playback to the transmitter. S10 controls the primary power relay on the remote power supply; a second set of contacts connects the receiver directly to

the monitor output when patch power is off. S11, the meter switch, was described above. A 28-volt pilot lamp is connected between V+ and V-, providing visual indication that both sides of the supply are on.

construction

The unit is built to mount in a station control console. The controls, microphone input, and monitor output are on the front panel. The back panel has a two-contact Jones plug for the telephone line, phono jacks for outputs to transmitter and tape recorder, a single pin jack for the push-to-talk lead to the transmitter, and an octal plug for the remaining connections, including ground.


Construction of the superior phone patch.

Amplifiers are mounted on two etched circuit boards, four on one and three on the other. These boards plug into 22-contact mounting strips attached to the bottom frame, which made it easy to remove the boards for experimental component changes during development. A metal strip across one end of the frame holds two transformers; the other, together with the H pad, is on a similar strip across the other end of the frame. Space was left on the frame for two more circuit-board mounts. One will be used to mount a VOX circuit that will control a vhf fm transmitter, and the other will permit installation of the three-unit biquad low-pass filter mentioned earlier if it proves to be desirable. The entire assembly was made from sheet aluminum and do-it-yourself strips and angles using home tools.

operation

For normal operation without a third party on the telephone, the TALK TO RADIO (S3) and MONITOR RECEIVER (S7) circuits are on. To set up a patch, S1 is turned on, the telephone circuit is activated, TALK TO PHONE (S2) and MONITOR PATCH (S6) are turned on, and S3 is turned off to keep the preliminary phone conversation off the air. Telephone contact is established using the station microphone and monitor. S7 may be turned off if signals from the receiver make it difficult to hear the party on the phone.

Contact between the parties is established by turning on the TELEPHONE RCVR (S4) and TELEPHONE XMTR (S5) switches, and proceeds with vox operation unless the telephone party trips the transmitter too often with laughter or interruptions. With S2 and S3 off, the operator can carry on another conversation or monitor another receiver without fear of interrupting the patch, but can monitor and break in if necessary.

I was tempted to title this article "an ideal phone patch" after several contacts, but the patch isn't – quite. The ideal patch would have constant level output to telephone and transmitter regardless of input source or signal strength. So far I have been unable to develop or find an agc circuit to accomplish this with operational amplifiers. I'd like to hear from the ham who has.

references

1. George P. Schleicher, W9NLT, "Phone Patching – Legitimately," *QST*, March, 1969, p. 11.
2. Motorola, Inc. Semiconductor Products Division publication "Motorola Integrated Circuits for MODEM and Terminal Systems" Section 4 – "Op Amps and Active Filter Design," July, 1970.
3. J. B. Berry, Jr., W4PME, "Legalize Your Phone Patch," *QST*, May, 1969, p. 17.

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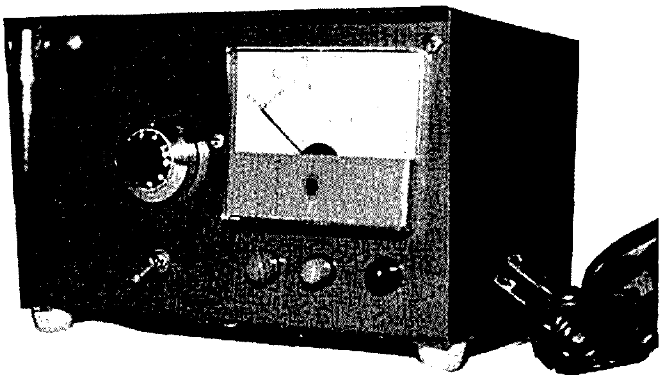


precision power supply

Looking for a laboratory-quality power source for your IC projects? Here's a likely candidate

Alan Nusbaum, W7SK, 3627 East Highland Avenue, Phoenix, Arizona 85018

After building many solid-state projects, I learned some interesting things about power sourcing, so I decided to build my own power supply. The little dandy shown on these pages is presented with the hope that others will benefit from my experience in this area.

First a "wish list" was made detailing everything the supply had to do to justify the investment of time and material. A axiom is in order at this point: Mor semiconductors are destroyed during ci cuit testing and trouble shooting tha expire from intrinsic device failure! Th principal cause of such destruction i over-voltage. So, heading the list is:

1. Variable voltage sourcing.

2. High-resolution voltage control with constant-current over the full range, and accurate resetability.

3. A high-resolution voltmeter, with 1% accuracy over the complete range.

4. Current-limiting control regardless of voltage setting.

5. Regulation better than .01 percent.

6. Ripple less than 50 mV.

7. A signalling method to identify which voltage range is being metered (a paramount requirement – a try at idiot-proofing!).

8. Sufficient reserve current to power the most ambitious solid-state projects.

9. Small size and the capability of operating continuously at full load.

The schematic of **fig. 1** shows the circuit.

the regulator

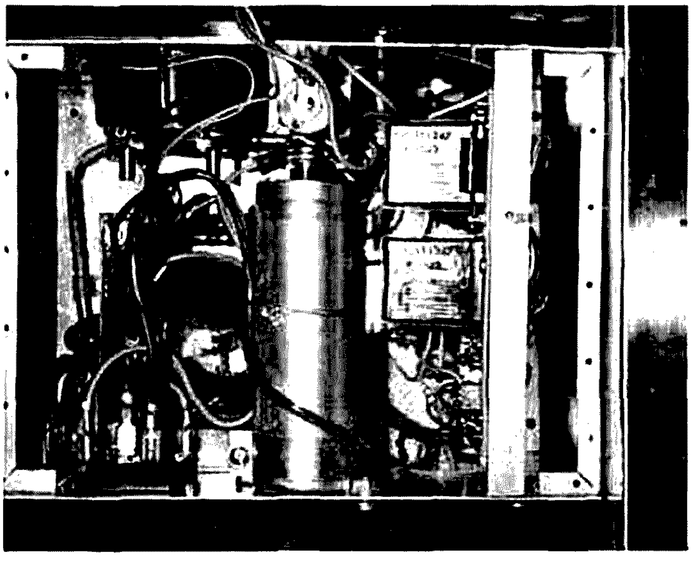
The Motorola MC1569R regulator IC was chosen because it satisfied the largest part of the regulation demands in one package. The MC1569R has a large power-handling capability, with a through-put current of 500 mA. Thus it was necessary to drive the low-beta series-pass transistor, Q1, to full saturation. The MC1569R will tolerate 36 volts input with safety. The circuit includes a voltage-control network consisting of a simple arrangement with a fixed 6.8k biasing resistor in combination with a 50k, 10-turn pot. This allows the output to be varied between 3.4 and 28 volts.

Still another attractive feature is the current-limiting ability, achieved by an external biasing transistor, Q2. Perhaps the most important feature is that the regulator won't oscillate (a plague affecting all high-gain op amp devices).

Two versions of this IC are available: the MC1469R with a rating of 35 volts maximum input; and the MC1569R, which has a maximum rating of 40 volts. If you choose a maximum of 34 volts to pin 3, the MC1469R will be more satisfactory.

The letter "R" denotes a TO66 package — do not use the "G" package (TO5 can) in this application, as its output is limited to 200 mA.

Underchassis view. The large 3300-uF, 45-Vdc electrolytic capacitor dominates the scene — essential for eliminating input transients to the regulator.



The variable-voltage control resistance (R1 in fig. 1) is determined by R1 = $(2V_O - 7)$ k ohms, which extrapolates to 53k ohms.* However, the nearest standard 10-turn pot is 50k, thus limiting the highest voltage to about 28 volts (minus the saturation voltage of the regulator transistor). Do not add any fixed resistance in series with this pot (R1), as it will increase the minimum voltage far above the useful range of RTL devices; i. e., 3.6 volts.

circuit description

Starting with the input to the series-pass regulator transistor (a Motorola 2N3771), the powering of all systems is done from 36 volts at the collector of this transistor. The output, or emitter, has an Ohmite 1.0-ohm pot (part no. 0101) in series with the load and a current-sensing transistor connected across this pot. This pot is set to sense 2 amperes. You'll note that an IR drop across this resistor upsets the V_{be} drop of the transistor (a 2N2219), causing the collector of this transistor to pull current from pin 4 of the MC1569R. This, in turn, drops the voltage output of the regulator, thereby biasing the 2N3771 base voltage down so that excessive current cannot be drawn—a Rube Goldberg action, but still a simple and effective means of device protection. A few words later about setting this potentiometer.

meter and lamp switching

In the 3.4- to 10-volt range, the green lamp is lit and the 0- to 10-volt scale of the meter is the relevant range. When the multiturn pot is rotated upwards, approaching 9.5 volts, zener D1 and its related voltage-offsetting diodes begin to conduct current via two paths. One is through D3 and D4 to ground; the other and lower offset path is through D2 and the Darlington pair transistors (Q4,Q5). At the point where the zener abruptly avalanches, the sum of all voltage drops through the second or lower offset path is

*Symbols used in this article are: V_O = output voltage; V_Z = avalanche (zener) voltage; V_F = forward voltage. *editor*

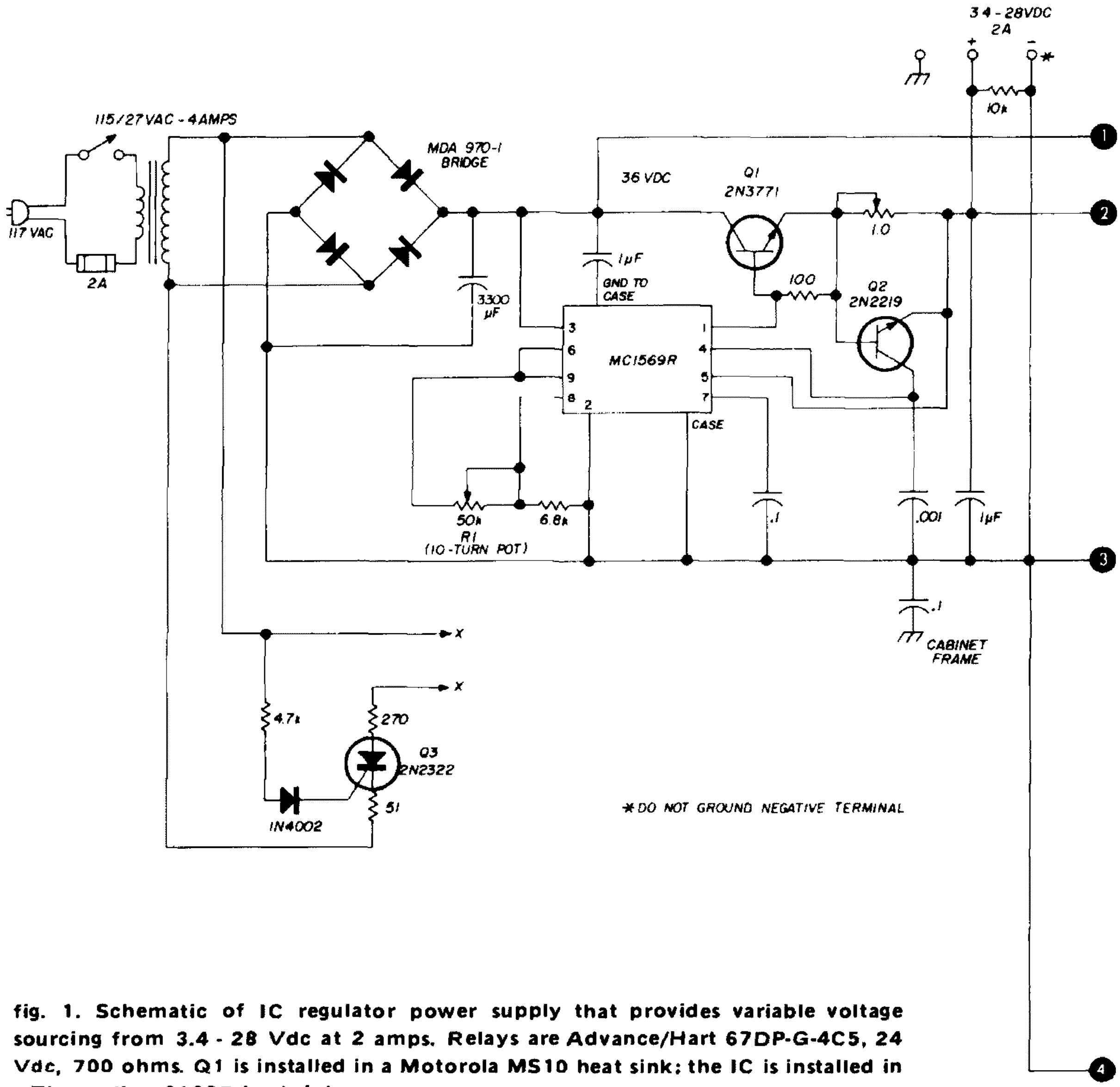


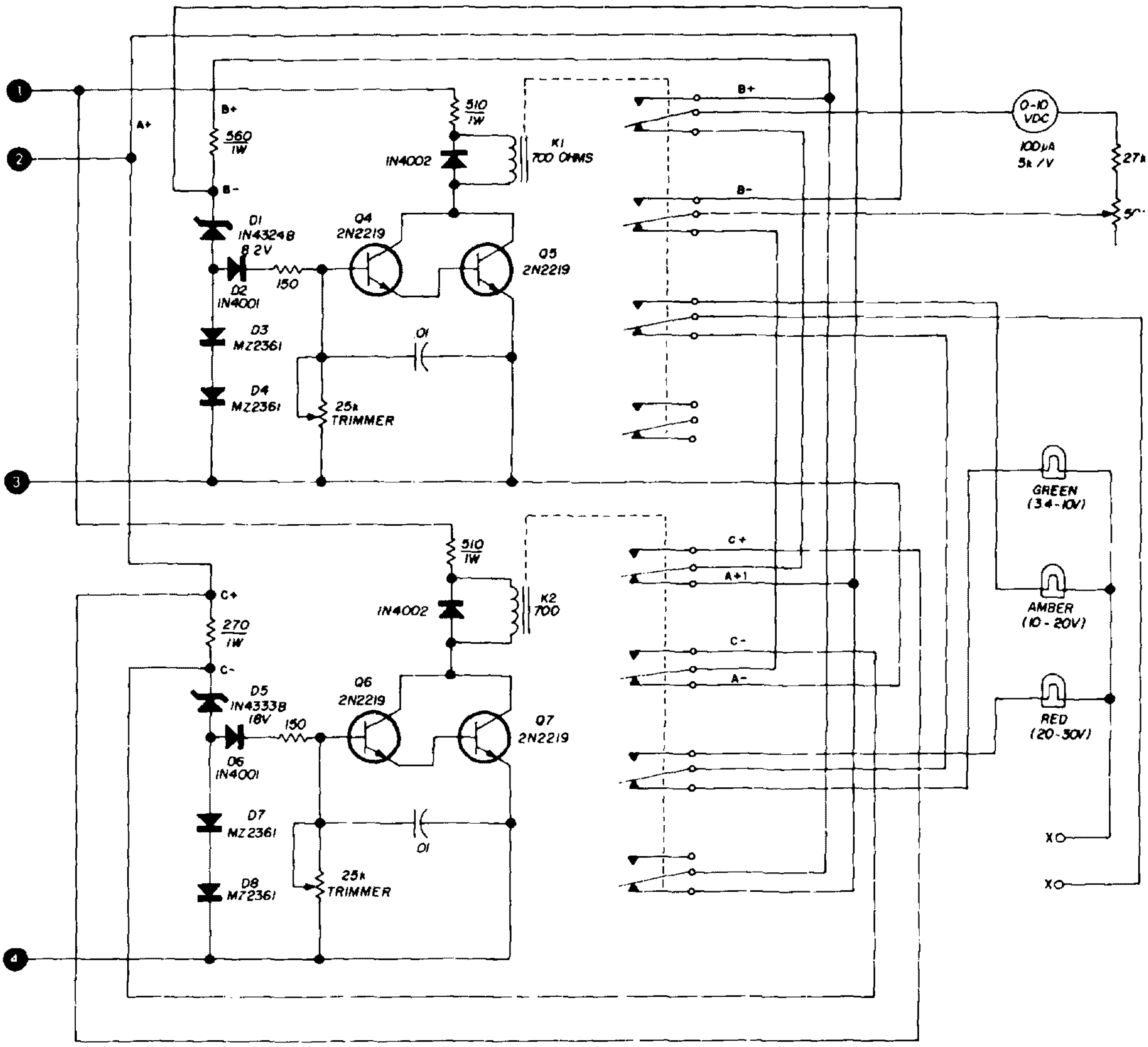
fig. 1. Schematic of IC regulator power supply that provides variable voltage sourcing from 3.4 - 28 Vdc at 2 amps. Relays are Advance/Hart 67DP-G-4C5, 24 Vdc, 700 ohms. Q1 is installed in a Motorola MS10 heat sink; the IC is installed in a Thermalloy 6168B heat sink.

equal to 10 volts. This network divides off excess current on each scale when the voltage is set near its upper range. Due to variations in beta of the 2N2219 (Darlingtons), the base biasing trimmer resistor may require adjusting.

A 25k trim pot in this circuit sets the proper toggle sensing of this network. When the 10-volt level actuates the Darlington-pair relay driver, the activated relay switches lamps from green to amber, and the metering shifts from 0-10 to 10-20 volts (full scale). This is accomplished by switching the negative side of the meter from ground to the cathode of the zener, so as to read the product of I^2R across the zener loading resistor on the 10-20 volt scale.

One may question the accuracy of such reading in view of the nonlinearity of most zeners. However, by using a 1-watt, 2% zener packaged in a metal can, the V_Z will remain stable provided the current is held within the maximum dissipation of the device. By not heating the zener, we avoid the effects of impedance change in the device. Meter comparison checks with a Simpson 270 showed no difference in scale accuracy.

As the voltage-control pot is again elevated, the next-level voltage shift energizes the second relay, disabling the



first sensing stage to conserve current. The indicator lamp shifts from amber to red, and the metering shifts from 10-20 volts to 20-30 volts (full scale), with readings taken across the second zener loading resistor. (Incidentally, this could continue ad infinitum.)

The meter has a 200-microampere movement at 5k Ω/volt. A fixed 10k series resistor plus a 50k trimpot sets the full scale to 10 volts. It reads all scales with this one setting – set and forget.

fine sensing control

You will note another trimpot in the base lead of the second Darlington switch. Again, variations in *betas plus* minute differences in V_F levels of the diodes dictate a fine tuning of the sensing network. Once this pot is set and the network shifts at 20 volts, the pot needs no further attention.

There will be some residual hysteresis when the voltage passes through switching levels. This is desirable, as it elim-

inates ambiguity in meter setting when voltages are set near the threshold of level change. The hysteresis is about 150 millivolts; but with regulation better than this figure, there will be no "hunting" when the supply is loaded and unloaded in fast sequence, such as with a pulsed load.

heat dissipation

The 2N3771 is used as a variable, electronically controlled rheostat, which in this application must dissipate considerable power in the form of heat. Let's consider a worst-case situation to show how much power we must sink off as heat.

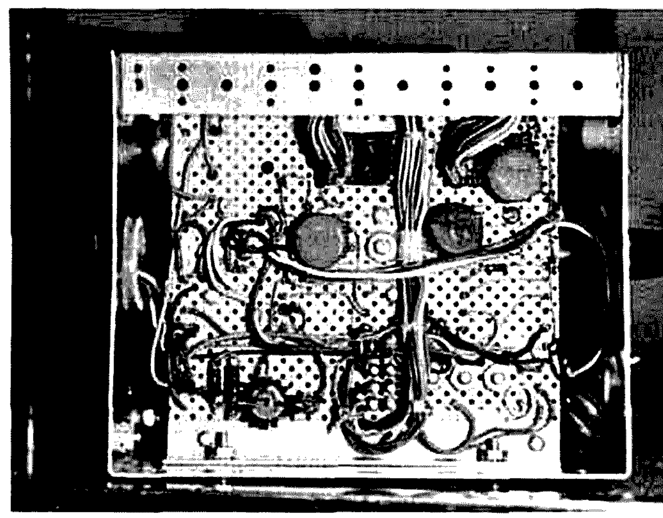
The input voltage to the 2N3771 is 36 volts from the rectifier-filter stage. The voltage control is set for 4 volts and we intend to use 2 amps from the supply. Therefore, we must drop 34 volts through the transistor, while it must pass 2 amperes or a total of 68 watts at 75° C. However, we're operating at 50% of the rated power load (150 watts at 25° C); but this is still a lot of heat to be cooked away – (try holding a 75 watt lamp bulb!).

One of the cardinal rules of transistor design is: "To prevent thermal damage, the heat extracted from the device must equal or exceed the heat generated." Therefore, an adequate heat sink for this transistor must be used, or else the safe operating area will be exceeded and the device will be instantly destroyed, even though it is running at 1/2 its rated power. In this application, a Motorola MS10 heat sink is used. Remember, it's the product of voltage drop and current that equals power-dissipation requirements!

This supply has been run on a heat test for 24 hours at 1/2 load without failure, although the cabinet was quite warm. A reading with a precision thermometer indicated nearly 75° C (167° F) at the edge of the sink – still safely within tolerance. Most tabletop projects don't require this amount of current, but it's nice to know you have a supply with a substantial reserve of power, just in case.

adjustment

To set the current limiting pot, first set the control, wired as a rheostat, at the 25% point or roughly 1/4 ohm. Then connect a 25-ohm, 100 watt adjustable bleeder resistor, set to 20 ohms, across the output terminals of the supply. With the voltage control set for 25 volts, slowly reset the bleeder to 13 ohms. Then increase the rheostat resistance to a point



Side view showing circuit-board components.

where the supply automatically switches to a lower voltage range, indicating current limiting.

Now reduce this control resistance until the voltage returns to 25 volts (at 13 ohms bleed) and quickly disconnect the load while watching the voltmeter. If the voltage shifts up or down, carefully fine-tune the current-limiting control while momentarily touching the load to the supply. Then when the full load-no-load voltage remains constant, this sensing control is in balance. Reduce the disconnected bleeder resistance to 10 ohms and attach the bleeder to the terminals as a check for supply shutdown, which should occur instantly. Lock the control shaft, and the supply is ready to operate.

I installed this control inside the cabinet, so that it can't be disturbed; it can, however, be mounted with the shaft extended and a knob installed for variable

current-limiting control from the front panel. Calibration and marking would be a simple task.

One last word about the MC1569R regulator. A small T066 heat sink is necessary to keep the temperature within bounds. A Thermalloy 6168B is suggested for this application.

construction notes

The supply was built in a 5 x 6 x 8-inch LMB aluminum box, with the circuit board mounted on the left side. There's nothing exotic about the assembly of the supply or wiring of relays and circuit-board devices. Number 22 teflon-covered wire (surplus) was used for these components. The high-current leads from the bridge rectifier to the pass transistor, then to the output binding posts, were no. 14 teflon-coated wire.

The output binding posts are spaced on 3/4-inch centers to accept Pomona-type ganged plugs. Three terminals are used, with the green post internally connected to the chassis, as both positive and negative leads are ungrounded.

conclusion

This supply has been a pleasure to use and has been indispensable in the construction of IC device projects — I'm sure it will be the same for you.

Those interested in further information in this application of ICs will find the material in the references very helpful. All are available upon written request from Technical Information Center, Motorola Semiconductor Products Inc., P. O. Box 20912, Phoenix, Arizona 85036. References 2 and 3 are contained in Motorola's "The Microelectronics Data Book," second edition, December 1969.

references

1. Motorola Application Note AN-457, "Switching Voltage Regulator Uses Discrete and Integrated-Circuit Approaches."
2. Motorola Application Note AN-473, "A Monolithic High-Power Series Voltage Regulator."
3. Motorola Application Note AN-480, "Regulators Using Operational Amplifiers."

ham radio

conversion from fast-scan to slow-scan television

Scan conversion offers an easy way to get on sstv — here are some ideas for sampling circuits and a multimode camera

R. F. Stone, W3EFG and A. B. Schechner, W3YZC

Slow-scan television (sstv) has been called one of the new and exciting frontiers of amateur radio, and rightly so. The number of experimentally inclined amateurs regularly transmitting and receiving good-quality still pictures with standard ssb equipment is steadily increasing. Listen on 14.23, 7.22, or 3.845 MHz, especially late in the evening, and at International sstv net time (2 p. m. EST) on 14.23 MHz to verify this for yourself.

First, however, you'll need an sstv monitor. Many construction articles on monitors have been published. One of the most recent describes an oscilloscope adapter for decoding sstv signals.[1]

What is needed to attract more amateur tv enthusiasts into the realm of sstv – for example, those now active on fast-scan atv in the 450-MHz band – is an easy way to use homemade or manufactured vidicon cameras on sstv.

For about a year we have been developing a simplified approach to scan conversion in addition to perfecting solid-state monitors. Don Miller, W9NTP; Ted Cohen, W4UMF; and others have pioneered with sampler cameras that convert fast-scan (FS) or standard tv signals to slow-scan (SS) standards (see reference 2).

W3EFG has built a multimode camera that includes a sampler board. It has performed well on the air and was first shown and demonstrated at the 1970 Dayton Hamvention. W3YZC started with a homebrew fast-scan vidicon camera and built a separate scan converter with power supply. This unit used the camera video output to derive sstv signals.

modifications for sstv

The circuits described in this article to obtain sstv signals can be easily adapted to almost any tube or transistor vidicon camera without destroying the capability of the camera for use elsewhere. The only change necessary in the camera is the addition of simple transistor circuitry to provide a switchable 15- or 60-Hz frame rate from the vidicon camera. This consists of adding a 4:1 unijunction-transistor countdown circuit to convert the 60-Hz pulse to 15 Hz and the addition of a larger value vertical-sweep capacitor. The circuit includes 3 transistors, a few resistors and capacitors, and a miniature dpdt toggle switch

***Co-author W3YZC joined the silent keys in January. Al's passing is a great loss to all who knew him.**

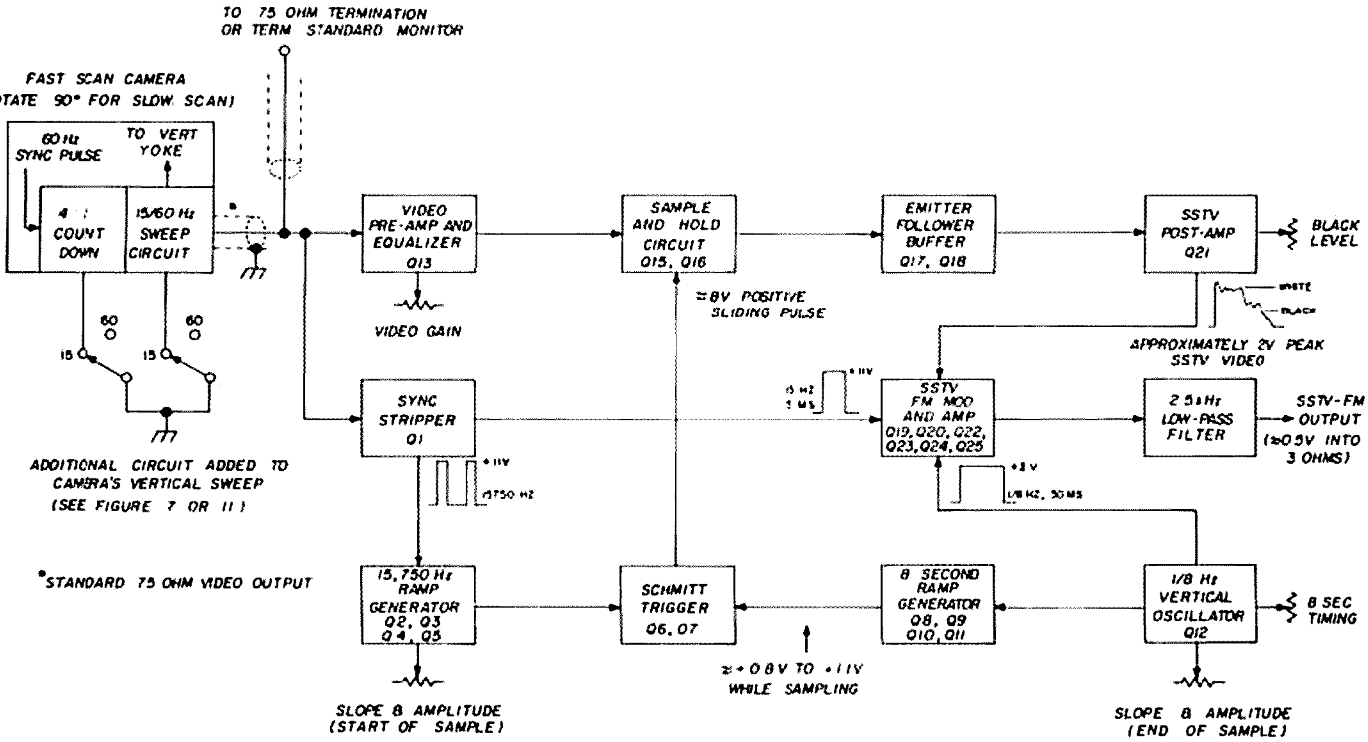


fig. 1. Block diagram of the FS-to-SS converter.

mounted on the back of the camera.

The composite video signal is used "as is" — the horizontal (15,750 Hz) and vertical (15 Hz) sync signals are stripped in the scan-conversion circuit. For slow scan, the modified FS camera must be tilted 90 degrees and operated on its left side, so that what was originally the top of the picture will be to the viewer's right. It's not necessary to reverse the vertical yoke connections because of the polarities used in producing the sliding pulse from the 15,750- and 1/8-Hz ramps (see reference 2).


Co-authors W3EFG . . .


and the late W3YZC via slow-scan tv.

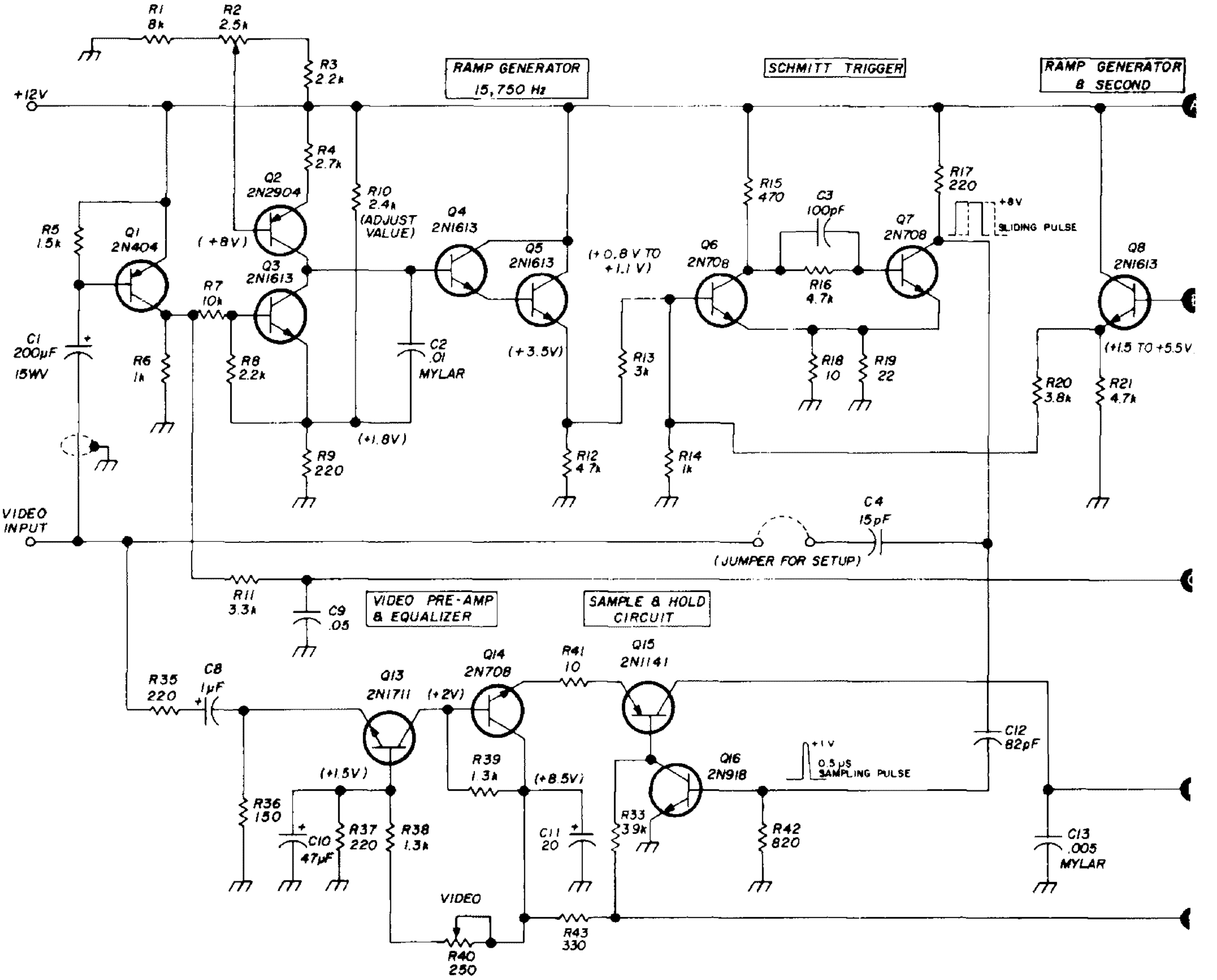


Fig. 2. Schematic of the slow-scan converter. Voltages and waveforms are shown for initial setup.

features

The scan converter design uses a minimum of components. Proven, debugged circuitry results in a scan converter that delivers a high-quality sstv picture with excellent resolution and contrast. The converter is easy to build and adjust. The accompanying photos, taken with the circuitry described here, pretty well speak for themselves.

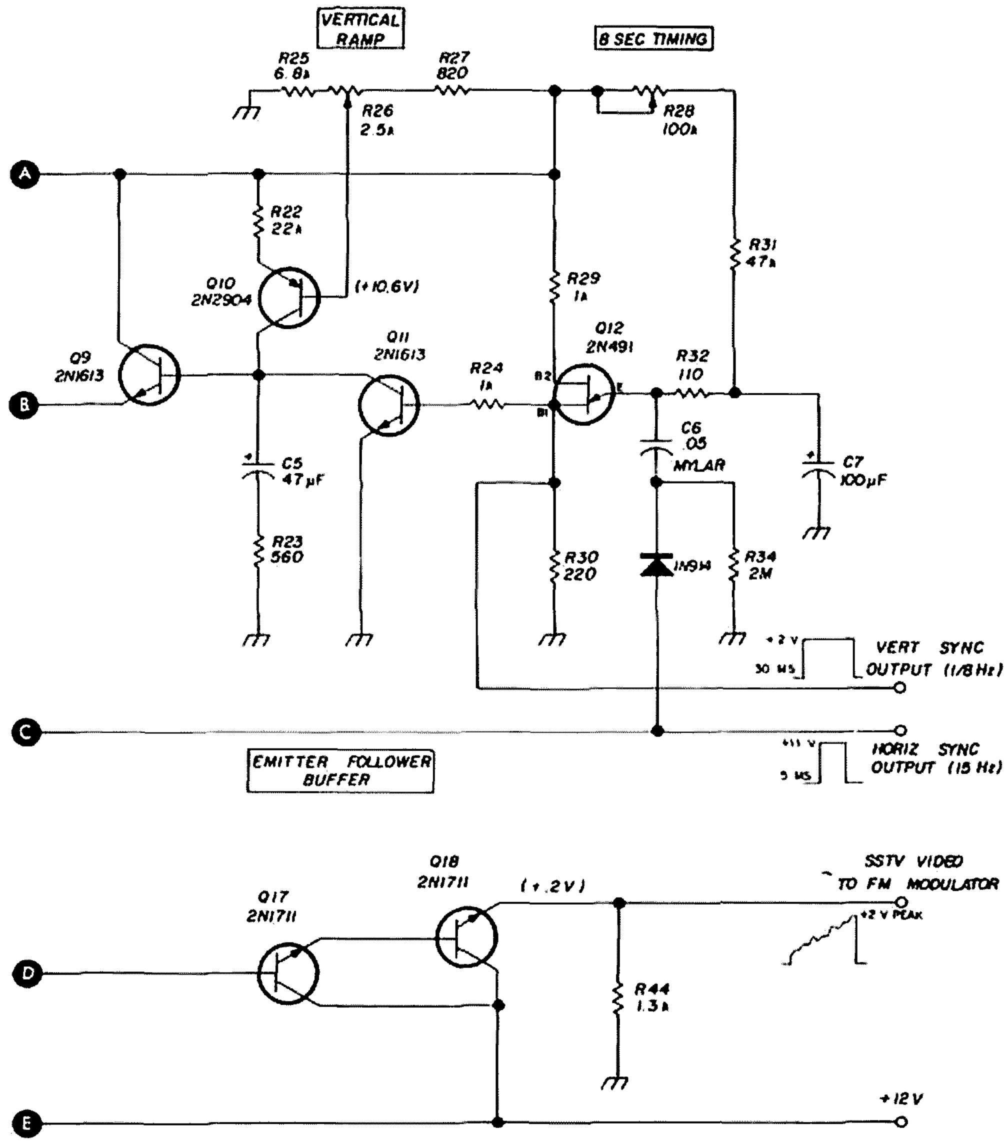
The following paragraphs described the basic scan converter circuits and give some ideas for packaging. We have also included a brief description of the multimode vidicon camera built by W3EFG, which includes design innovations from circuits by W3YZC and from the articles on homebrew cameras [3], [4].

The multimode camera was the result of an attempt to build a densely packaged, universal camera later used on atv, sstv and that could be adapted for experimentation on vhf and uhf with pseudo-random narrow-band tv having motion capability.[5]

A printed-circuit board and parts kit for the scan converter is available from HAL Devices, Box 365H, Urbana, Illinois 61801.

converter functional description

The functions of the scan converter are depicted in **fig. 1**, which shows interconnections for a standard vidicon camera. The 4:1 countdown from 60 to 15 Hz and the sweep-circuit modification will vary somewhat from camera to



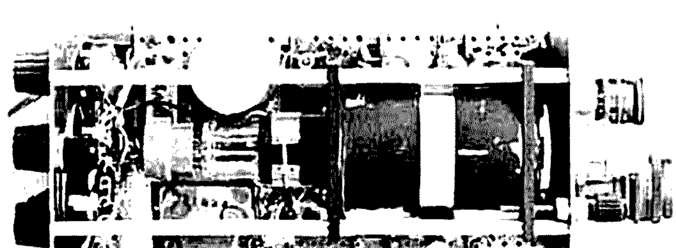
camera. However, circuit details associated with a typical modification may be deduced from the vertical-sweep circuit of the multimode camera described later (see fig. 7).

The block diagram shows the various set-up and operational potentiometers as well as typical waveforms and voltage levels. If positive-going horizontal and vertical sync pulses are available in the camera, sync stripper Q1 may be eliminated. This was done in W3EFG's camera, since the FS board was directly beneath the SS converter and fm modulator board.

scan circuit

The scan converter schematic is shown in fig. 2. Composite standard video is fed to both the sync stripper, Q1, and to the video preamp and equalizer, Q13. The 15,750-Hz pulses initiate a linear 15,750-Hz sawtooth voltage that is added to a 1/8-Hz linear sawtooth voltage from an 8-second vertical oscillator, Q12. Note that Q12 is a free-running oscillator, which is synchronized with the start of the sstv horizontal line by the network

W3YZC's homebrew FS vidicon camera.



composed of 1N914 diode, R34, and C6.

The two ramps are added across R14 of the Schmitt trigger, which causes the circuit to fire over the range of approximately 0.8 to 1.1 volt. This produces a sliding pulse at the collector of Q7 that is differentiated by C12 and R42 to give a 0.5-μs pulse (approximately). The 0.5-μs pulse opens gate Q16, which samples the video through Q15 to holding capacitor C13. The charge on C13, which is the sstv sampled video, is coupled by a Darlington-pair buffer, Q17, Q18, to the modulator.

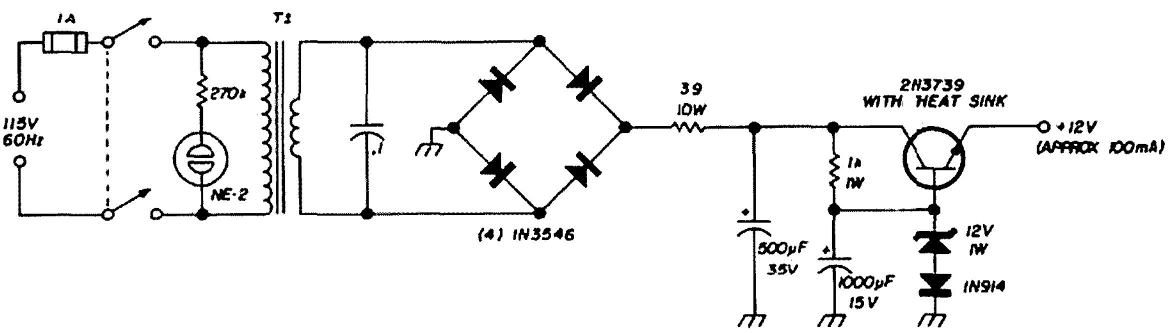
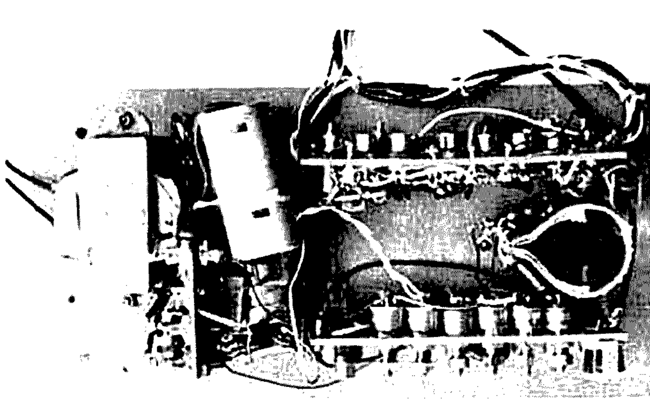


fig. 3. Suggested power supply for the sampling converter and fm modulator. T1 is a 24-26 volt, 1 amp filament transformer (Stancor P6469, Lafayette 99E62663, or equivalent).

The value of holding capacitor C13 was optimized to produce the best sstv picture. Emitter followers Q17, Q18 provide a high-impedance source to the collector of Q15.

A common-base-connected video preamp, Q13, drives the sample-and-hold circuit. The incoming standard FS video is fed through an equalizing network, R35, C8, R36. The time constant of C8, R36 was chosen to attenuate the lower frequencies of the fast-scan video. We found this to be necessary with most vidicon cameras, because the vertical shading showed up as a bright or dark area on the left or right side of the sstv picture.


Slow-scan converter and power supply used with W3YZC's camera.

The video pot, R40, adjusts the bias of Q13 and controls both gain and the sync clipping level. The gain of Q13 is approximately equal to the ratio of R39 to R36 and may be adjusted if required.

device selection

Special mention should be made about the characteristics of the sampling-gate transistor, Q15. Considerable investigation preceded the selection of the optimum device to produce the best resolution consistent with good gating action.

A transistor with poor gating characteristics will produce a leak-through pattern on the sstv monitor, resulting in streaks as the picture is painted across the tube. Best operation was obtained from a high-speed germanium switching transistor – the 2N1141. The 2N1141 has been around for quite awhile and is still listed in the catalogs. A lower-frequency 2N1142 was tried, but it produced slightly degraded performance.

All transistors shown in the schematics have worked well. Countless other npn and pnp devices with similar or related characteristics are available and may be used in these circuits. For example, the inexpensive epoxy transistors have been used in most circuits and work fine. The

2N3906 will probably replace the 2N2904, and the 2N3904 will likely work in place of the 2N1613 or the 2N697.* We suggest that if you wish to use junk-box transistors, stick with the silicon types, with the exception of the 2N1141 for Q15.

power supply

The sampling system is totally intolerant of 60- or 120-Hz hum in the video. Even a few hundredths of a volt ripple will show up as vertical bright bars, slight undulations, or wavy kinks on the horizontal sweep in the sstv picture. For this reason it's essential that the power supply be well filtered, with less than 0.01-volt ripple or better. The circuit of fig. 4 is recommended for the scan converter and modulator.

In some vidicon cameras, stray magnetic fields may occur at a 15-Hz frame rate. Thus it may be necessary to disconnect the primary of a self-contained power transformer. Filament and vidicon voltages can then be supplied from an external supply placed at least three feet from the camera. Only operating experience will reveal this requirement.

*See Motorola's "Semiconductor Data Book," 5th edition, October 1970, for recommended replacements and substitutions for these devices. editor.

slow-scan modulator

An sstv fm modulator similar to that used by most slow scanners is shown in fig. 3. Its setup on frequency is straightforward and has been previously described.[2] Voltages and input waveforms are shown in fig. 1. It is desirable to first apply 10 volts to T. P. 1 and set the 2300-Hz white frequency, then ground T. P. 1 for the 1200-Hz sync frequency from the multivibrator. This should be

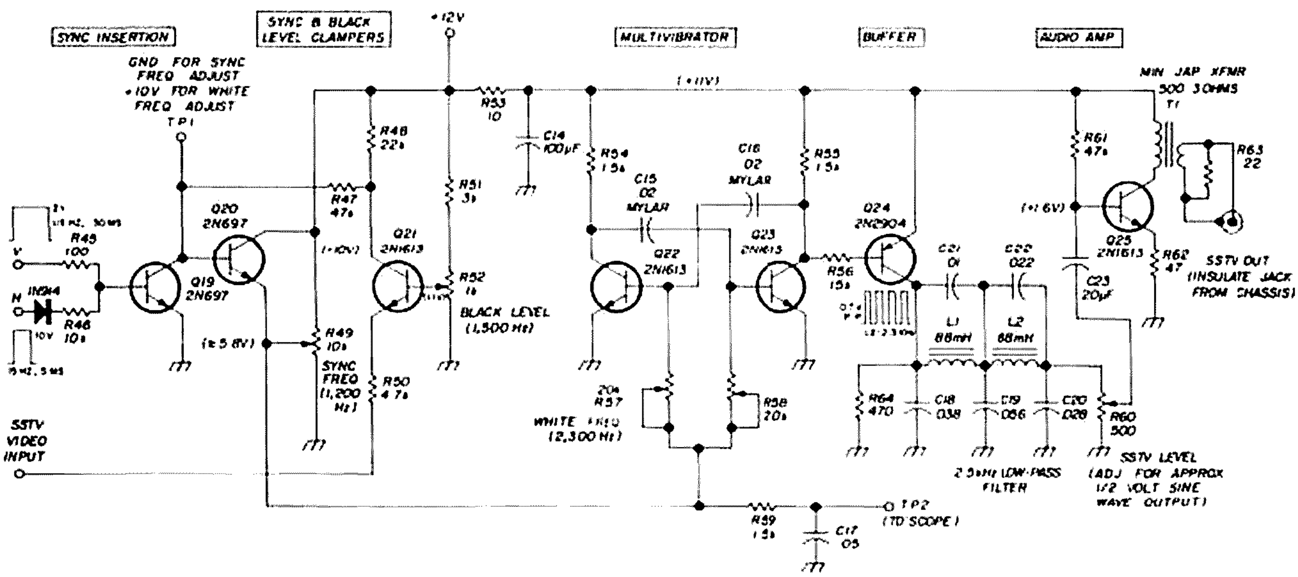


fig. 4. Modulator schematic for FS-to-SS converter.

Identification chart used by W3EFG.

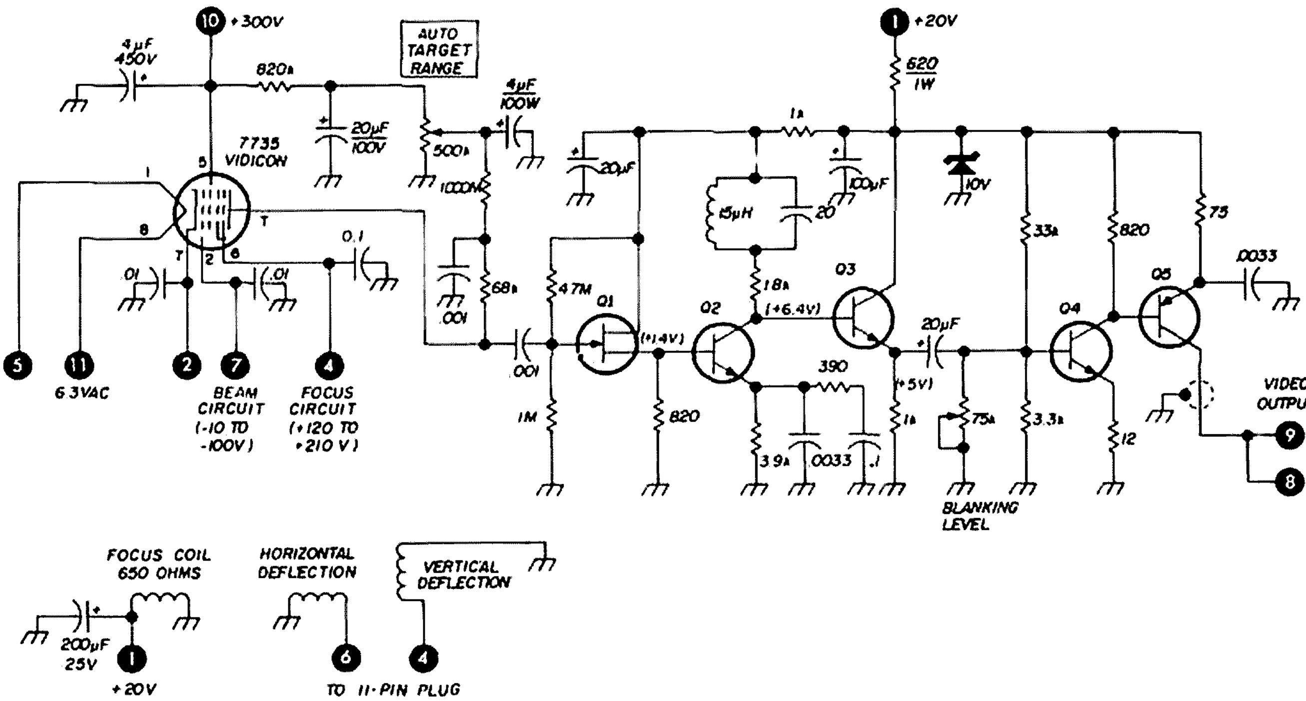
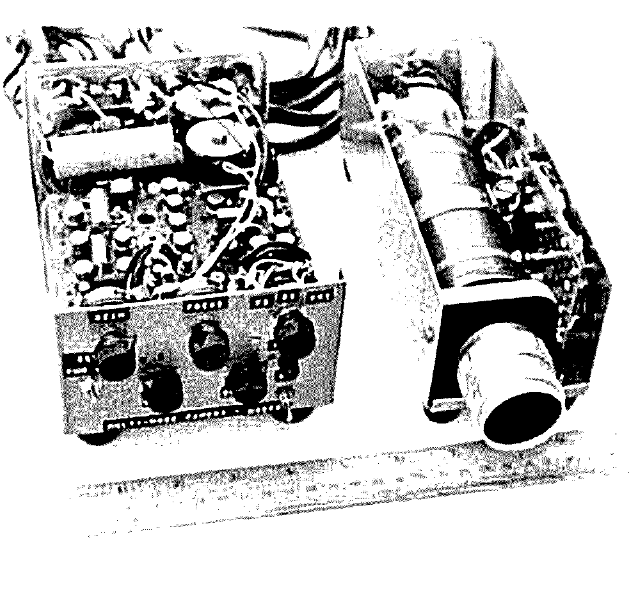


fig. 5. Multimode vidicon camera head with video amplifier.

repeated several times. T. P. 2 (filtered by R59 and C17) is a good point to connect an oscilloscope for monitoring black and white level in the scene being televised. Using the lens iris and video pot R40, adjustments should be optimized to prevent white clipping and to obtain proper black-level excursion.

Following the multivibrator, Q24 provides drive to the 2.5-k Hz low-pass filter. L1 and L2 are 88 mH toroids, tuned by the capacitor values shown. The odd

Multimode vidicon camera built by W3EFG.

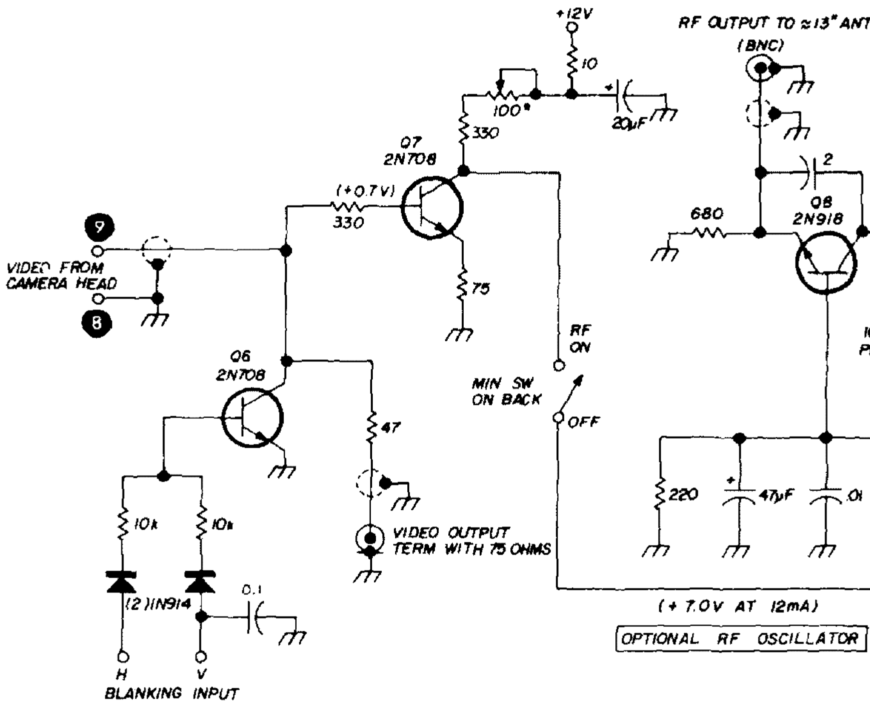


values should be obtained by combining standard-value mylar or mica capacitors.

The filter may be mounted on a small board that can be tucked into an available space, along with the power supply. Note that sstv output is at a low impedance and at a fairly high level (approximately 0.5 volt across 3 ohms), which is compatible with 3-8 ohm outputs from receivers and recorders. The sstv output is taken from a phono jack insulated from the case with fiber washers.

set up procedures

It is convenient to view the FS video on a fstv video monitor when setting up the start of the sampling pulse and the point on the horizontal ramp where the pulse ends and returns to start a new sstv frame. Since Q13 tends to isolate feedthrough of this pulse to the video input, C4 should be jumpered to the video input, for initial setup. The differentiated spike will then be clearly visible on a regular monitor as a thin, bright vertical line. Its width is indicative of the sampling-pulse width, and its excursion from the viewer's right on the screen to the viewer's left indicates that portion of the FS picture being sampled.



*ADJUST FOR OPTIMUM SYNC LEVEL ON RF SIGNAL
L1 = 3-1/2 TURNS NO. 16, 1/4" ID, TUNES CH 7-13 WITH 10pF PISTON CAPACITOR

fig. 6. Video blanking mixer and video-modulated rf oscillator for the multimode vidicon camera.

A 1:1 aspect ratio is used in sstv, so a narrow portion to the left and right of the FS picture will be lost. Since there is some interaction in adjustments, a careful procedure must be followed to center the sampling range while watching the FS monitor. (Two small pieces of black tape on the monitor screen will help.)

The horizontal-ramp pot, R2, will affect the beginning of the sampling pulse. The sampling speed, and hence the limit or end of a frame is determined by the vertical-ramp pot, R26. Resistor R10 sets a dc level of 1.8 volts, which may be adjusted to shift the sampling-position range of R2 and R26.

A vernier range of adjustments is possible for R2 and R26; however, the final setting of both controls is quite critical. Screwdriver-adjusted potentio-

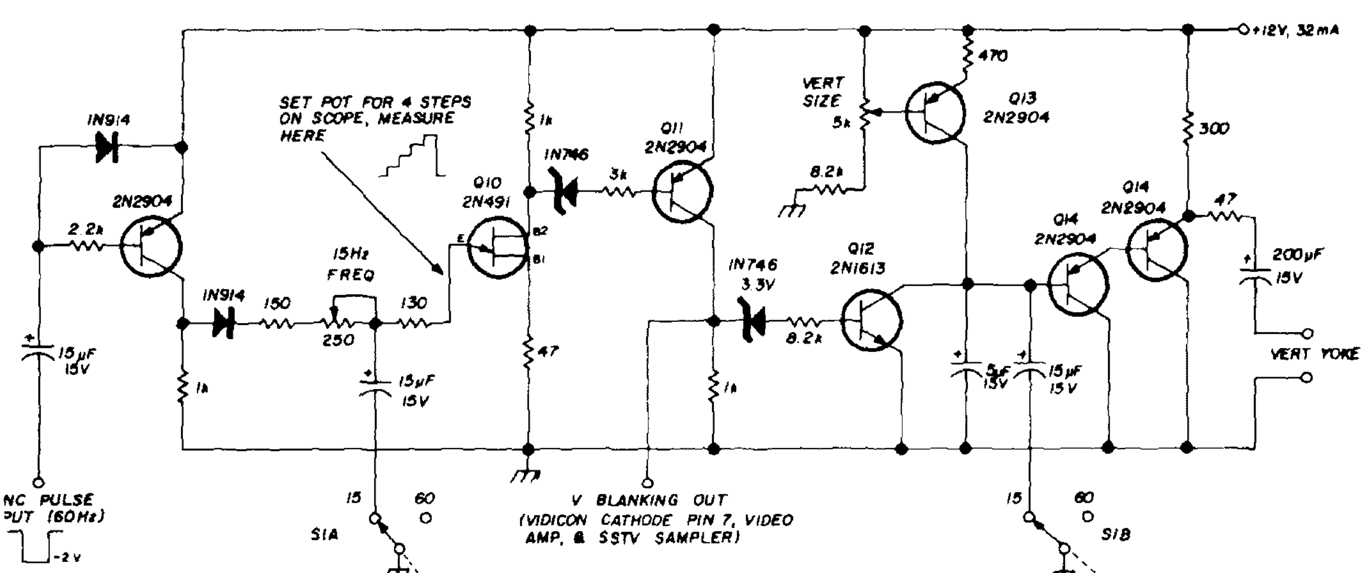


fig. 7. Vertical-sweep circuit for FS or SS camera.

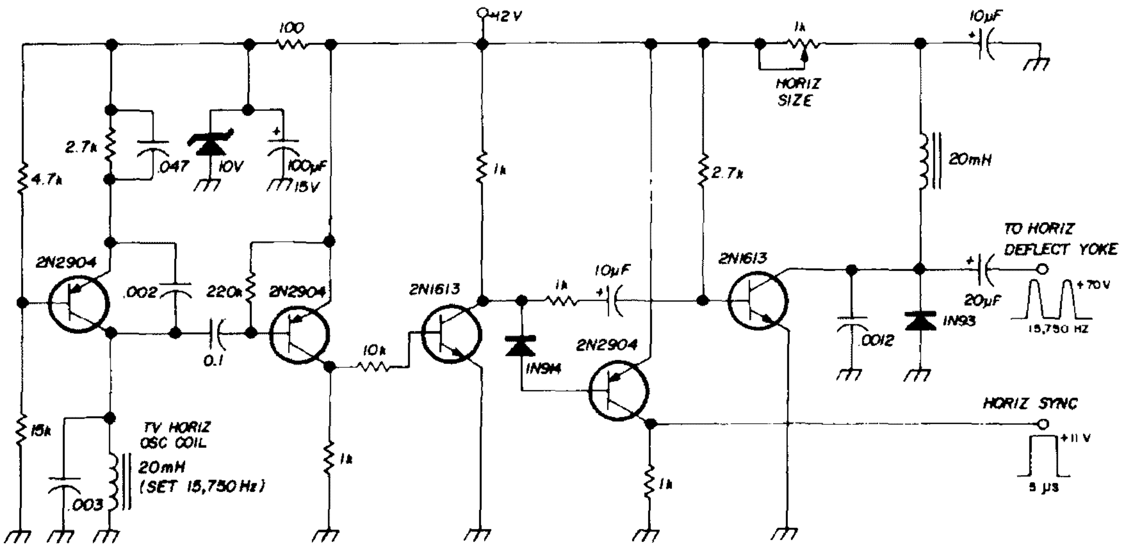


fig. 8. Horizontal-sweep circuit for FS or SS camera.

meters with locking nuts are recommended.

The voltages and waveforms shown in fig. 2 are provided as a guide. These are approximately equivalent to those obtained by the authors. Patience during adjustment and optimization of values may be required.

packaging

The converter shown in the photo is contained in a 10 x 5 x 3-inch minibox which provides ample room for two small power transformers and associated switches, pots, and connectors.

The vector-board chassis could be adapted to a printed-circuit board. Perhaps some enterprising slow-scan enthusiast with PC-board layout facilities will tackle this job and make such boards available. This would certainly give sstv

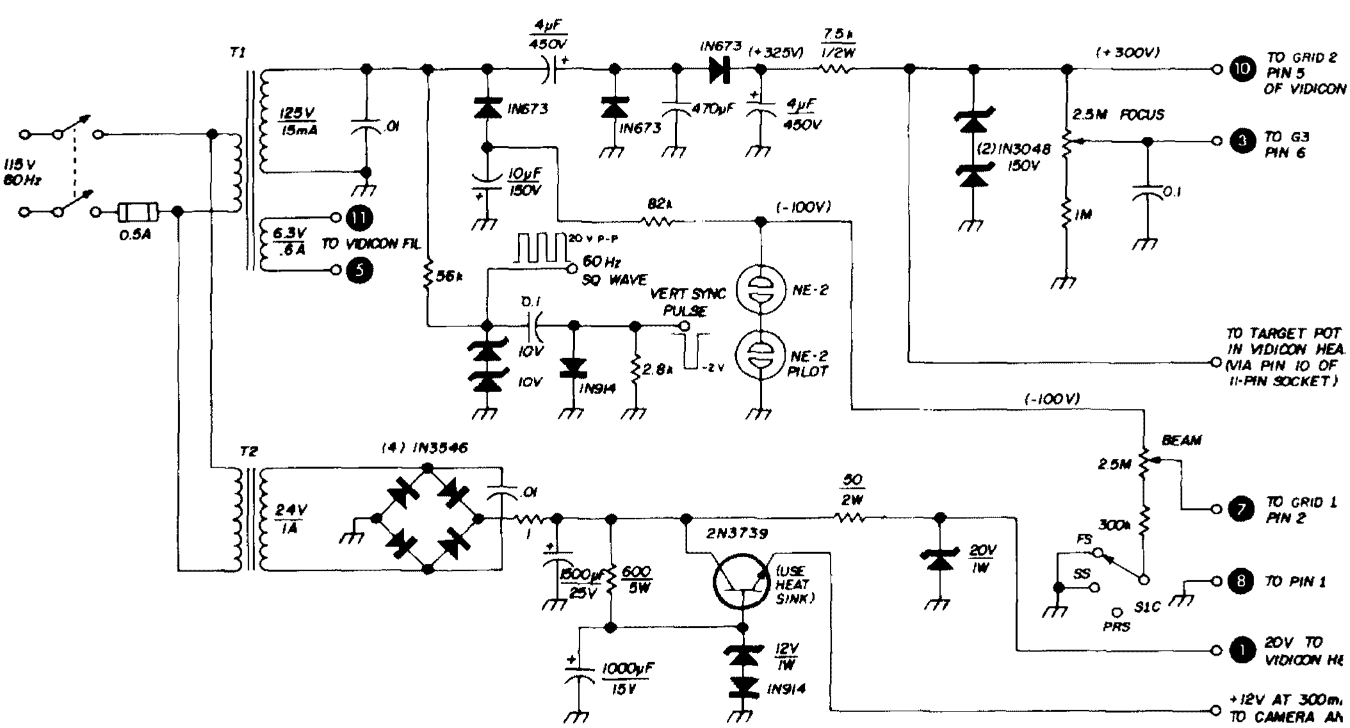


fig. 9. Suggested power supply for FS or SS camera. T1, Lafayette 33E34059; T2, Lafayette 99E62663.

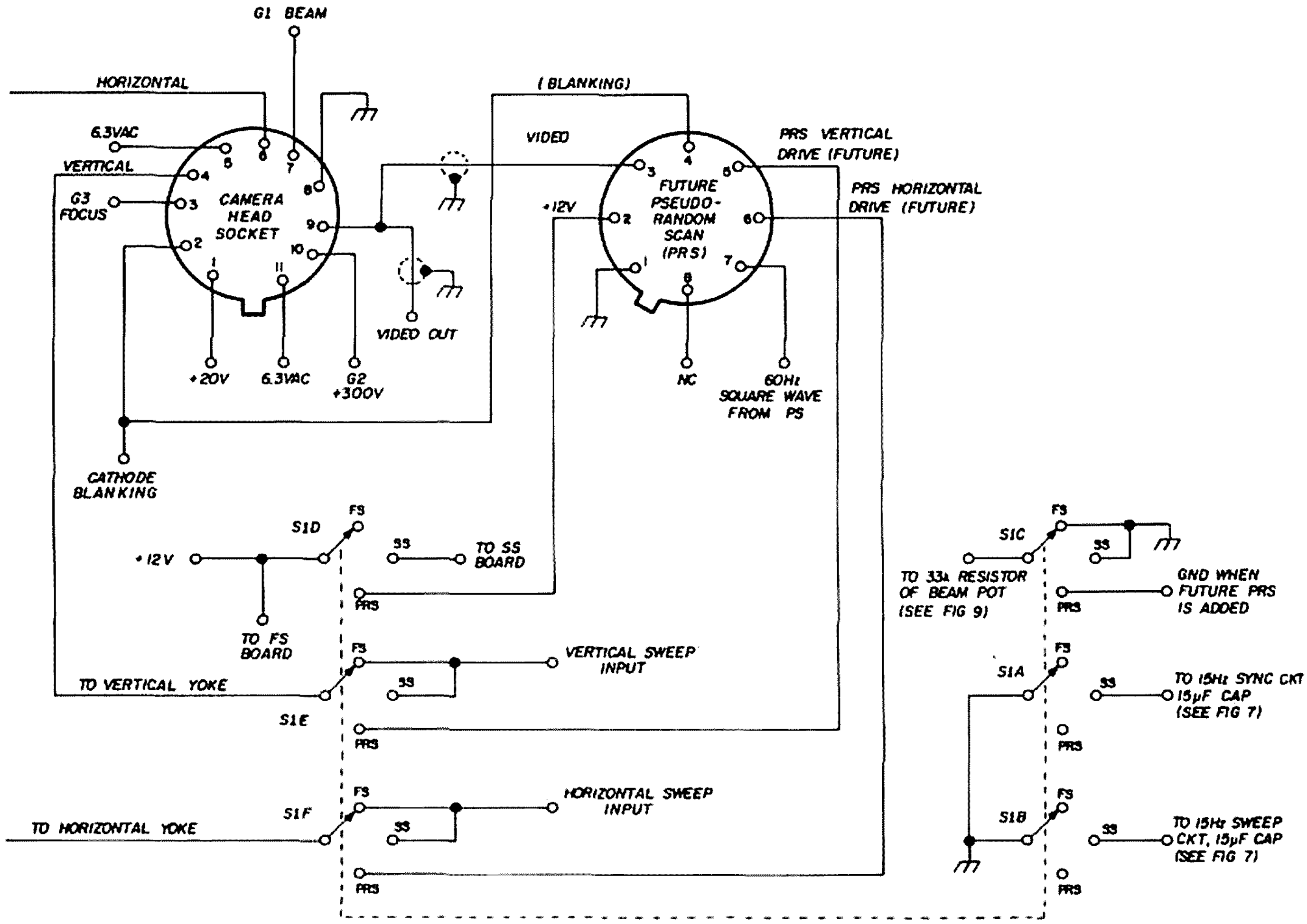


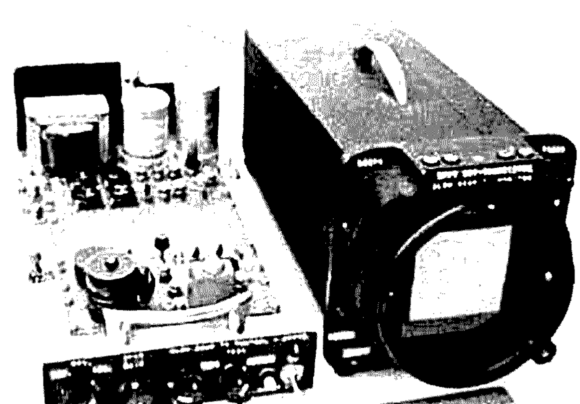
fig. 10. FS/SS/PRS switching circuit for the multimode camera.

activity a further boost with more sampler cameras on the air.

Scan converter board wiring is not critical as to lead lengths, interconnections, or placement of wires and components. Standard construction techniques were used. Push-through vector board pins were used for tie points. Resistors and capacitors were wired point-to-point to the transistor sockets. Several straight buswire runs on pins for the 12 V and ground leads were used. Because of the high frequency components in the narrow sampling-pulse circuitry, single-point ground returns for Q6, Q7, Q16, and Q15 are recommended. Also a short run of 75-ohm mini-coax from the back panel video connector to the video input is essential, plus a short length of coax across the board to C1 of the sync stripper transistor, Q1. These coax cables should be grounded at only one point on the board to prevent ground currents in the FS video.

fast-scan camera

The circuit details shown in figs. 5 through 10 are provided with a minimum of explanation for those ambitious slow-scanners who may wish to duplicate a



Solid-state sstv monitor using a 5FP7 tube. The photos of the ID chart and the authors were made from W3EFG's camera and this unit.

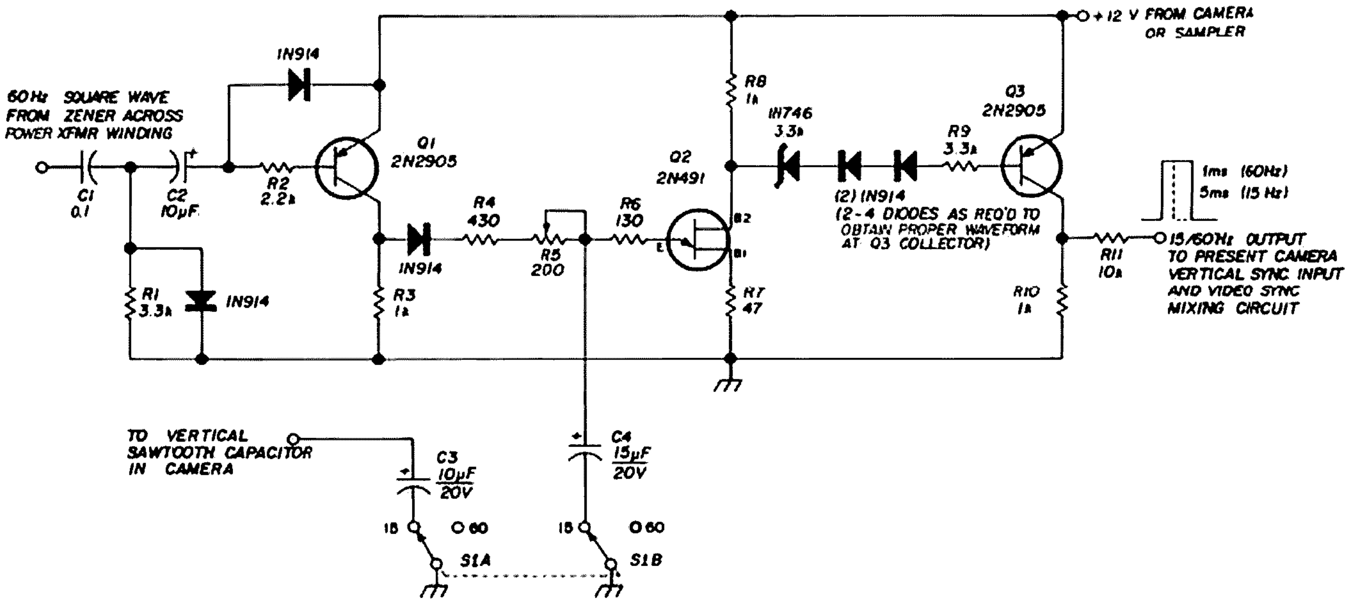
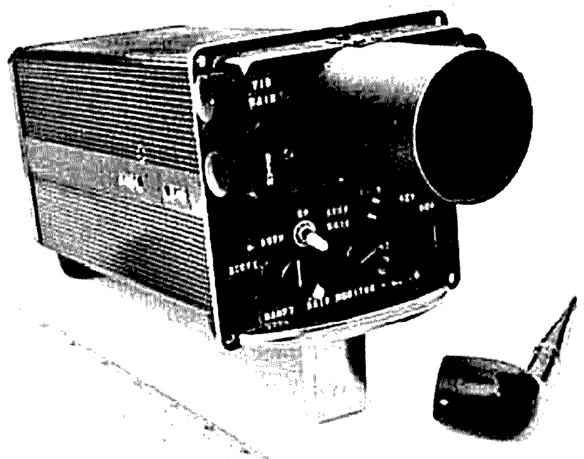


fig. 11. 15/60-Hz vertical sweep modification for a standard vidicon camera for use on sstv. C1, R1, C2, R2 determine the 60-Hz pulse width. The sstv 15-Hz sync pulse width is determined by R6.

typical homebrew multimode camera. These circuits, which are used in W3EFG's camera, are generally straightforward and can be modified to drive vidicon focus – deflection-coil yoke assemblies having different impedances.

With the exception of the video board



Another "seeing aid" used by W3EFG. Monitor features solid-state circuitry built into an old Waterman scope case. It provides a 1¼-inch-square image, which is magnified to 2 x 2 inches with a lens attachment. Inquiries from serious hams on this and the other equipment described are welcome. Send your questions and comments, together with a self-addressed stamped envelope, to W3EFG, 506 Eastwood Drive, Exton, Pennsylvania 19341.

and vidicon yoke in the 3 x 2-½ x 8-inch camera head, all circuitry in figs. 6 through 8 are on a vector board measuring 4-5/8 x 3-11/16 x 3/32 inches. This FS board is mounted in the 5 x 7 x 3-inch minibox beneath the SS converter board. Switching is shown in fig. 10 for the various operating modes.

A suggested power supply for the entire camera is shown in fig. 9. Only the portion of the supply associated with the 24-volt transformer need be built if only the scan-converter board is constructed. However, the supply can be built to supply total power to an existing camera if hum problems develop, as mentioned earlier.

vertical sweep

Attention is called to fig. 7, which shows the vertical-sweep circuit. To make the 15 to 60 Hz frame-frequency modification to an existing camera, as mentioned earlier and shown in fig. 1, parts of the circuit in fig. 5 can be used. We suggest that the components associated with Q9, Q10, Q11 be put on a small vector board, with the indicated switching circuit (see fig. 11). This board may then be placed in any available space within, or external to, an existing camera. Careful study of the particular camera circuit will show the correct tie-in points.

This modification should work equally well with random interlace and 2:1 interlaced cameras.*

conclusion

These fast-scan camera circuits are third- and fourth-generation versions used by the authors. All are proven circuits and produce excellent FS and SS pictures.

This article is intended to provide active slow-scanners and atvers an optimum means of producing high-quality, live sstv pictures. You may wonder if "holding still" for eight seconds is difficult. Our experience shows this is no problem. Since we're scan-converting or sampling a fast-scan camera, there is no latent change pattern on the target, such as may occur with the open-shuttered vidicons usually employed in the SS mode.

Other advantages of the sampling approach include the ability, in real time, to focus the vidicon optically and electrically. Second-order benefits include automatic light compensation via automatic target circuitry.

We would like to thank all sstv amateurs contacted for their continued interest in our activities and for their encouragement. In particular, thanks are due W9NTP and W4UMF for their inspiration and pioneering efforts to promote the sampling technique, which appears to be a much-needed break-through for sstv amateurs.

references

1. Bill Briles, W7ABW, and Robert Gervenack W7FEN, "Slow-Scan TV Viewing Adapter for Oscilloscopes," *QST*, June, 1970.
2. Don C. Miller, W9NTP, "Slow-Scan Television," *CQ*, Part 1, July, 1969; Part 2, August, 1969.
3. R. O. Gibson, W8TYY, "The Vidicon Minicamera," *73*, April, 1968.
4. Reed Fisher, W2CQH, "Transistor TV Camera," *A5* Magazine, May, 1969.
5. Sid Deutsch and R. Simpson, WA2PAX, "A Continuous Motion Narrow Band Television System," *CQ*, April, May, December, 1968.

*A G. E. 4TE23 camera was so modified with excellent results.

ham radio

six-meter antenna coupler

This 500-watt antenna coupler for 50 MHz matches both balanced and unbalanced feedlines and uses commercially-available hardware

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Current six-meter transceiver and transverter designs follow the general supposition that the operator will be using 50-ohm coaxial cable and an antenna system of like impedance. Output tank circuits are usually pi-networks designed to match a 40- to 60-ohm load and allow little deviation beyond these limits. On the other hand, the amateur buys 50-ohm coax, assembles a 50-ohm antenna to the manufacturer's specifications, and assumes the load presented to the output of the transceiver is 50 ohms. In most cases it is not unless careful attention is given to swr measurements, antenna match adjustments, antenna location and a host of other variables.

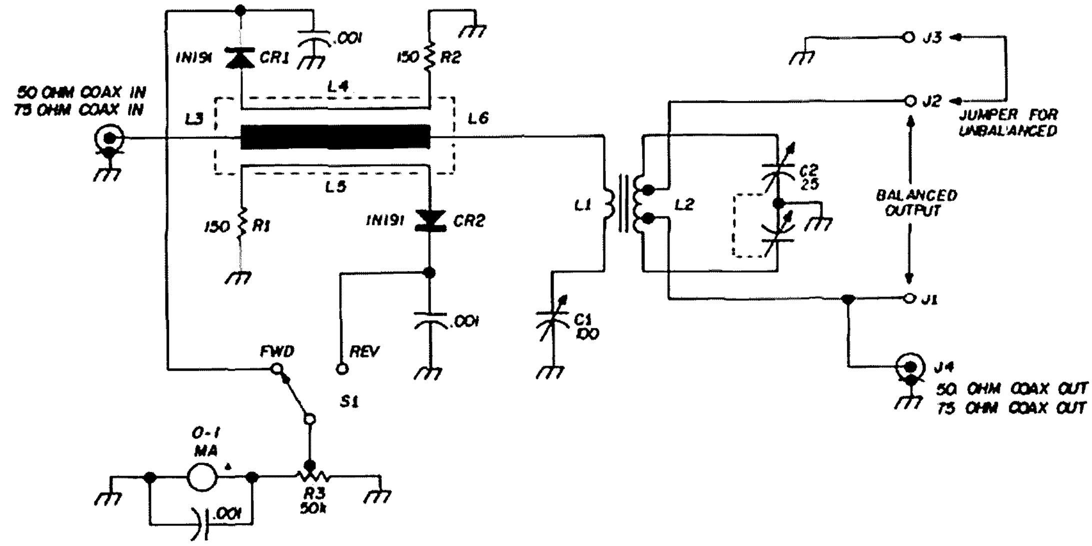
To compound the problem amateurs begin loading the system with goodies such as swr bridges, monitor scopes and the like which do nothing to improve transmission line flatness, and in most cases add to the mismatched condition by adding impedance "bumps" to which the transceiver has difficulty accommodating. The end result is a transceiver which is difficult to tune and erratic to operate.

In these situations the often maligned and usually misunderstood antenna coupler can be used to its best advantage. Many hams who lack experience with couplers have the mistaken idea that they are a panacea solution to antenna problems. However, an antenna coupler will *not* compensate for a misadjusted antenna or one in need of repair; nor will it compensate for final-tank-circuit inefficiencies.

It *will* provide the transmitter with a constant load impedance irrespective of antenna variations. In addition, it will assist in receiving by attenuating spurious signals through the introduction of another tuned circuit in the input.

construction

Many articles have been written on antenna couplers. Unfortunately, these designs often require the builder to fabricate coaxial cavities and various plastic



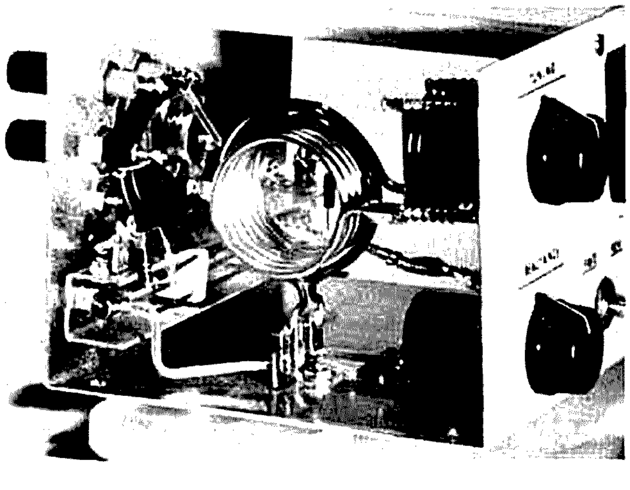
C1	25 = pF butterfly capacitor (E. F. Johnson 167-22)	L4, L5	0.062 brass rod, 4" long (Heathkit 40-99)
C2	100 pF variable (Hammarlund HF-100)	L6	coaxial section (Heathkit 40-100)
L1	2 turns no. 14 enameled, 2-1/8" diameter, over center of L2	J1, J2, J3	5-way binding post
L2	7 turns no. 10 spaced ½ wire diameter, 1½" diameter, tapped 2½ turns from each end	R3	50k pot, linear taper (Heathkit 10-11)
L3	0.250 brass rod, 5" long (Heathkit 40-98)	S1	spdt toggle or 2-position rotary (Heath-kit 63-177)

Swr bridge uses coaxial line section from Heathkit bridge.

spacers and other hardware to incorporate an swr bridge into the coupler. The coupler described here takes the lazy man's approach and uses commercially-made hardware throughout; it requires an absolute minimum of fabrication.

The coupler consists of a conventional input/output circuit of a 2-turn input coil tuned with a 100-pF variable to ground. The input coil surrounds the output coil and is inductively coupled; the output inductor consists of 7 turns of no. 10 copper wire tuned by a 7-35 pF butterfly variable to ground. The output inductor is tapped for 300-ohm balanced output at 2½ turns from each end.

Swr bridge circuitry is mounted on rear panel of chassis.



swr bridge

An swr bridge is included to provide a self-contained functional unit. The heart of the bridge consists of a coaxial cavity line and rf pickoff elements from a Heathkit swr bridge. The coaxial-section components were puchased from Heath for less than $2 and eliminated the need for sheet-metal bending, parts fabrication and chasing all over for plastic, brass rod and the like. The meter is a Radio Shack Micronta, 0-1 mA, catalog no. 22-018. The meter face was removed, and the original numbers and lettering were

erased with a coarse typewriter eraser. The scale was relabeled to correspond with swr. Lettering was done with rub-on letters. The meter face was then sprayed with clear acrylic. A full-scale drawing of the meter face is shown in **fig. 1**.

assembly

Drilling templates are shown in **figs. 2** and **3**. The chassis should be drilled and all holes deburred. Burnish all exterior surfaces with 00 or 0 grade steel wool. Wipe all surfaces to be painted with a degreasing solvent or white vinegar to remove surface dirt and oils. The unit in the photograph was built for K1ZFE and was painted flat black and Dove Gray to match his equipment. Krylon spray colors were used; excellent results were obtained by applying several light coats with plenty of drying time in between. After painting, each cabinet half was baked in an oven for one hour at 200° to assure complete drying. Black transfer lettering was added to the front panel and Dymo labels on the rear.

Mechanical assembly should begin with the swr coaxial section and SO-239 input connector. The section must be shortened by one inch to fit this cabinet. The output end of the coax section center conductor is mounted to a 6-32 ceramic standoff insulator. Then add the terminal strips, terminating resistors, rectifier diodes and .001 bypass capacitors; assemble the switch, sensitivity pot and all interconnecting wiring. Wiring from the rectifier diodes to the switch should use shielded wire.

The two front-panel air-variable capacitors and the terminal tie point on the left side (bottom) of the chassis should be asembled to the cabinet next.

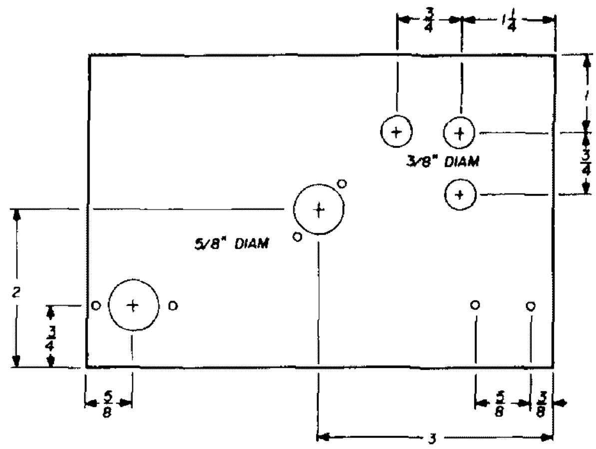


fig. 2. Rear-panel layout for the coupler.

Tie the ground lugs of each air variable together and to the center ground terminal of the tiepoint; use no. 14 copper wire. *L2 is connected to the butterfly* capacitor by two copper sleeves made by enlarging the diameter of a ½-inch long piece of 3/16-inch copper tubing. Use a 100- or 150-watt soldering iron for soldering.

fig. 1. New meter face for Micronta meter.

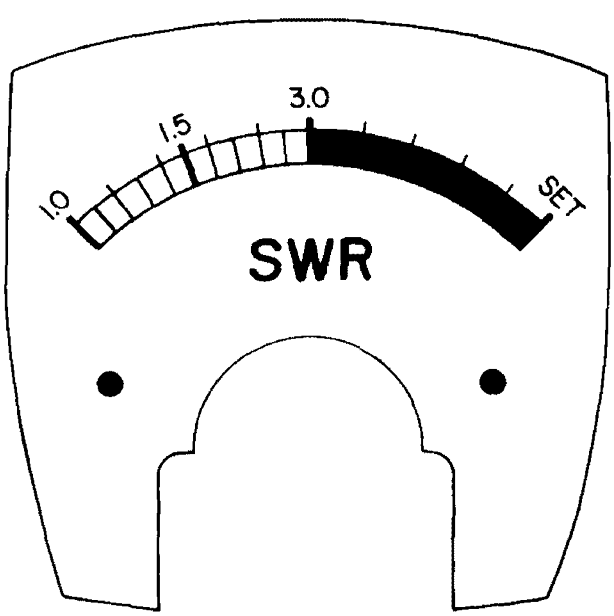


table 1. Parts list for 6-meter antenna coupler.

qty	description	cost
1	Chassis Box (LMB 564 N)	$2.95
1	Meter (Radio Shack 22-018)	2.95
1	Butterfly variable (E. F. Johnson 167-22)	3.45
1	Hammerlund variable (HF-100)	2.65
1	Coaxial section (Heath 40-100)	.20
1	Rf driver element (Heath 40-98)	.55
2	Rf pickup element (Heath 40-99)	.20
3	Plastic spacer (Heath 255-12)	.30
3	2-lug terminal strip (Heath 431-14)	.30
2	SO-239 coax connectors	1.70
1	50k pot (Heath 10-11)	.50
1	2-position rotary switch (Heath 63-177)	.85
2	1N191 Diodes	.60
2	.001 disc capacitors	.40
2	150 = ohm resistors, ½ watt	.20
1	Ceramic standoff	.45
3	Binding posts (Superior)	.90

Be sure to slide L1 over L2 prior to completing the assembly of L2 to the butterfly. Bend the legs of L1 to fit the spacing of the two ungrounded tie points and make sure that the spacing of L1 around L2 is even. Connect the output of the coaxial line to one end of L1; connect the other end of L1 to the stator of the 100-pF variable.

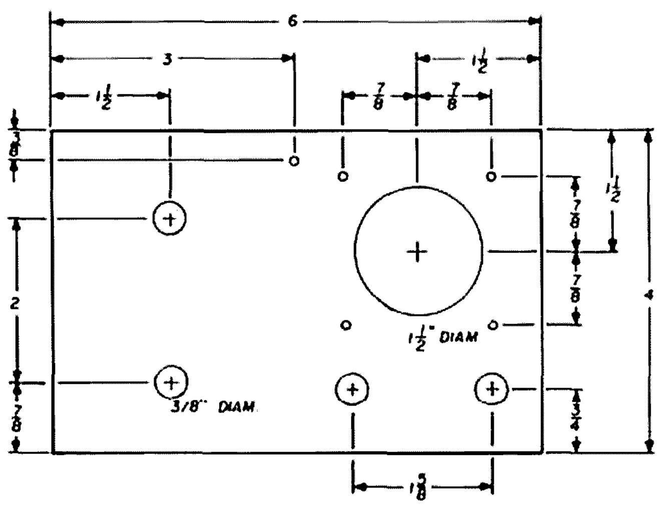


fig. 3. Front-panel layout for the coupler.

Assemble the 5-way binding posts to the chassis and the output coaxial connector to its mounting hole. Connect the upper two binding posts to the taps on L2 with a short piece of 300-ohm transmission line (preferably foam filled). Connect the output coax connector to the 5-way binding post nearest the center of the chassis. The lower binding post nearest the outside edge of the chassis should be connected to ground. Assemble the meter to the cabinet and wire it to the 50k pot. All interconnecting wiring in the L1/L2, C1/C2 section is done with no. 14 wire.

test and operation

If the coupler is to be used from one coaxial cable to another, install a jumper wire between the grounded binding post and the one directly above it. This *must* be done to use the coupler for coax-to-coax transfer. Connect the coupler between the transmitter and the antenna or a 50-ohm non-reactive dummy load such as the Heath-kit cantenna. Tune the transmitter for low power output. Set the swr bridge switch to forward and adjust the sensitivity pot for full meter deflection. Reset the switch to reverse and adjust C1 and C2 for minimum reflected power. It should be possible to adjust the reflected power to zero.

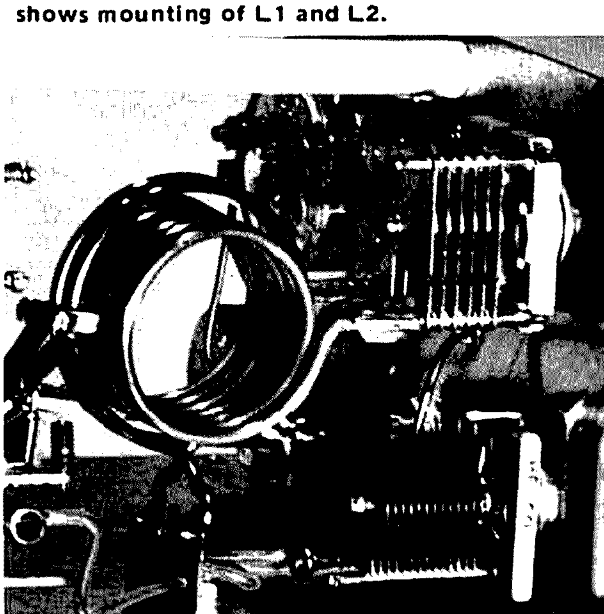
When the coupler is adjusted for minimum reflected power the transmitter can be tuned for full output power. It is possible to move 100 kHz either side of the tuned frequency before the coupler must be re-adjusted. Once the antenna coupler is set for minimum reflected power, leave it alone — tune the transmitter final to obtain maximum output, not the coupler.

conclusion

This unit will handle up to 500 watts without problem. Insertion loss has been measured at 1 to 2 percent. However, the improved efficiency provided by the coupler more than compensates for the low insertion loss. For those amateurs who already have an swr bridge, the coupler section could be built into a smaller cabinet and used with the existing bridge. Tuning procedures would be the same. Just be sure to install the swr bridge between transmitter output and the coupler input.

ham radio

Inside view of antenna coupler shows mounting of L1 and L2.



versatile vox

Solid-state circuit for voice-operated transmission and fast cw break-in

Richard Bain, W9KIT, 4915 Ridgedale, Fort Wayne, Indiana 46815

Presented in this article is a solid-state vox unit designed as a companion to my 9-MHz ssb generator described in an earlier issue of *ham radio.*[1] I had three objectives in this design:

1. Operation with all normal settings of the ssb generator audio-gain control and with all settings of the hang-time control.

2. Operation in the cw mode, with a minimum of components, and with fast pick-up time to avoid lost characters.

3. Fast break-in operation.

circuit description

The circuit that fulfills these requirements appears in **fig. 1.** The function switch is shown in the ssb vox position. The vox input from the generator feeds the base of Q1 through CR1. Note that a positive input is required. To provide this, the ssb generator should be modified as shown in **fig. 2.** The 47k resistor is connected to ground to quickly discharge the vox filter capacitor.

With no vox input, Q1 has no drive; consequently Q2 and Q3 are turned off. When a positive input arrives from the generator, Q1 turns on Q2, which in turn drives Q3 into saturation, causing relays K1 and K2 to operate. One contact of K1 grounds the transmitter keyline and opens the mute line to the receiver. The second contact may be used to close relays in a linear amplifier when the vox relays close. The hang time is primarily determined by the C3, R4 time constant. The value of R4 was selected so that several times more current would flow through it as into the base of Q2. This minimizes the effect of Q2's base current on the hang time and consequently minimizes temperature effects on the hang time due to transistor gain variations.

The network of CR2, CR3 and R6 insures proper operation of the vox circuit at very short hang time settings. Without these components, C3 must charge and discharge through a major portion of R4's resistance. This means that if the hang-time control is set for a half second as an example, several dots or dashes of a letter must be sent before the vox relays will hold in. With the network shown, C3 charges rapidly to nearly the emitter voltage of Q2 through R6 and CR3 upon closing the key. Upon releasing the key, Q2 receives base drive through CR2, thus avoiding the high attenuation

path through R4. The quick-charge network also helps prevent relay chattering with short hang-time settings in the ssb vox mode. At the maximum hang-time setting, CR2 is shorted out and CR3 is back-biased with the key down, so neither affects circuit operation. CR3 is also back-biased when the key is released regardless of the hang-time setting and thus has no effect on the discharge curve of the hang-time network.

The 1k resistor in series with the base of Q3 limits the base current to safe values, in this case about 8 mA maximum. The diode across the relay coils clips the large voltage spike created when Q3 turns off in a short time period, such as during break-in operation.

In the cw vox position of the function switch, the keyline is connected to the key jack. Resistors R1, R2 and R3 form a summing network between the positive vox supply and the negative keyline voltage. R2 is adjusted so that the voltage at the base of Q1 is zero. The resistance presented to the keyline is close to a megohm, and so should not disturb the bias circuits of most transmitters. The voltage at my key is -35 volts. For higher keyline voltages, increase the size of R3, and reduce it for lower voltages.

cathode keying

You may have a cathode-keyed transmitter with which you may wish to use this circuit. In that case, Q1 may be changed to a pnp transistor and components added to the transmitter as shown in **fig. 3.** The voltage of the zener diode must be larger than the vox circuit supply voltage and must also be large enough to completely cut off the final tubes. The values of R1, R2, and R3 must be large enough to hold Q1 base current to a safe value. Values as shown in **fig. 1** are satisfactory. R2 is adjusted so that Q1 base voltage just equals the emitter voltage. Note that the vox input now comes into the base of Q2. This means that a lower-impedance source of this vox voltage is desirable for rapid charging of the hang-time network. The R-ZD1 combina-

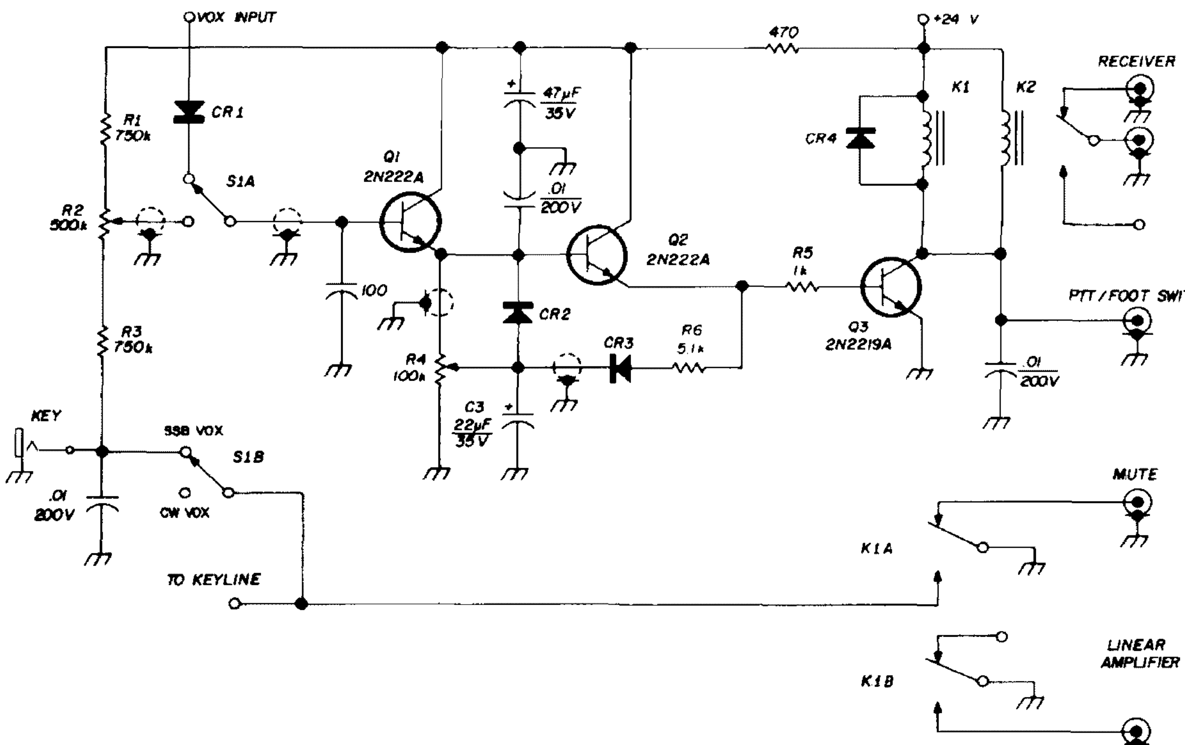


fig. 1. Vox schematic. Fast cw break-in is also featured.

tion could be replaced by a voltage divider if regulated voltage is available in the transmitter.

break-in

Break-in operation in the cw vox position is possible by turning the hang-time control to minimum. The two relays (if they are capable) will now follow the key. For break-in operation up to several hundred watts, the vacuum relay shown is recommended for its fast operation and ability to hot-switch several amps. K1, the other relay in my unit, is a small two-pole sealed 28-volt relay. I have tried other types, and the surplus 28-volt leaf-type relays will follow a key at moderate speeds. The extra noise created can be damped with felt between the relay and the chassis. Reed relays may of course be used, and some of the heavier ones may handle moderate powers.

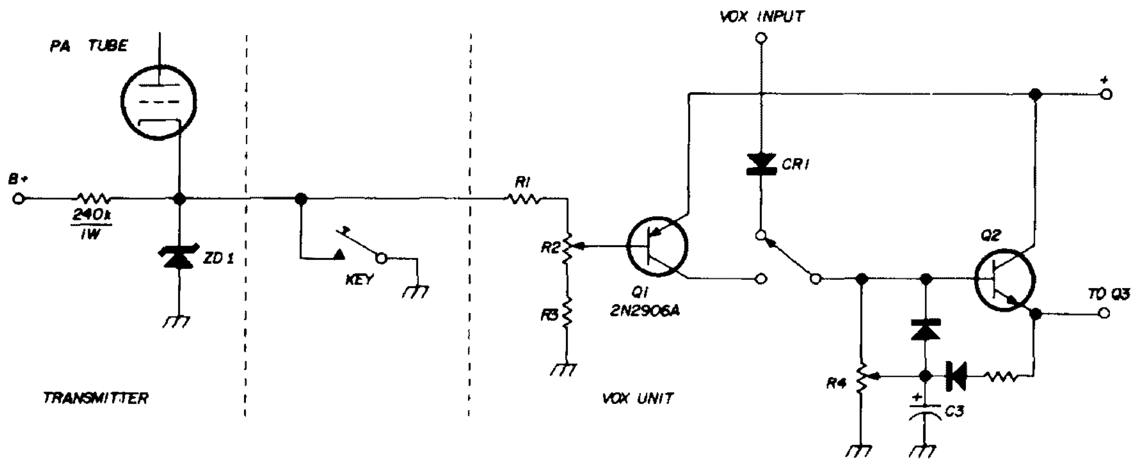
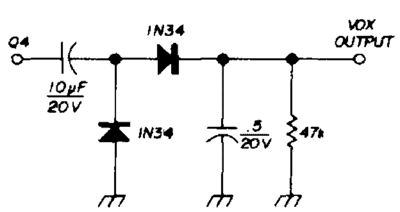


fig. 3. Vox modifications for use with cathode-keyed rigs. Value of resistor in tube cathode circuit is for B+ = 300 volts. Zener-diode voltage is 60 volts dc at 1 mA.



2. Modifications necessary to author's ssb generator described in ham radio, December 1970 for use with this vox circuit.

device considerations

The transistors are on the expensive side, but satisfactory substitutes can be found. As an example, Poly Paks lists the 2N2222 at five for a dollar. While their V_{ce} is 30 Vdc, they should work well if care is taken to hold the supply voltage to 24 Vdc or less. The power dissipation of Q1 and Q2 is negligible, but that of Q3 depends on the relay current required. As an example, with the key down, Q3 is saturated with 120 mA through it in my vox unit, but the voltage across it is less than a volt. The power dissipated is less than 120 mW. When the key is released, however, the voltage across the transistor rises, and the current drops. At the point where 12 volts are across the transistor, 60 mA may flow through it for a short-term dissipation of 720 mW. The average dissipation, however, would be under 500 mW, and a 2N2222 could be used for Q3, but with little margin. At this point, a 2N2218A would be a good substitute (and far cheaper) for the 2N2219A.

If you wish to use whatever you can dredge up from the junk box, try the following measures to compensate for lower-beta transistors: decrease the values of R1, R2, R3 and R5. Operation with 12-volt relays requiring more current than those listed may also require a decrease in these resistor values.

construction

My unit was built on a 1-1/2 x 2-inch piece of perf board, which was mounted on the rear of my transmitter chassis. R4, C3 and S1 are on the transmitter front panel, and R2 is on the chassis rear apron beside the perf board. Shielded wires to these controls are a must to avoid rf feedback. The antenna relay should be located remotely from the rest of the circuitry, and leads going to it should be well bypassed.

other uses

The applications of this circuit are numerous. It could be added to a transmitter to eliminate throwing of switches to go from transmit to receive, or to control a linear you may have added to your station. It could be set up for battery operation with low-current relays and used on Field Day to mate someone's transmitter with someone else's receiver. It could also be used to key a tape recorder to record slow-scan tv or someone's voice. Or, how about recording passing satellite signals? Whatever your use, happy voxing!

reference

1. R. Bain, W9KIT, "A Filter-type SSB Generator," *ham radio*, December, 1970, p. 6.

ham radio

"All right buster! Back to your radio and never mind the skirt selectivity."

how to design a frequency calibrator for your receiver

Complete design procedure for a novel marker generator consisting of a constant-current source, unijunction oscillator and tuned output amplifier

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Recent changes in amateur band-edge assignments for various license classes, and lack of crystal calibrators in many popular transceivers, has resulted in a rash of magazine articles on crystal calibrators. Some circuits are better than others and make provisions for 25-, 50- or 100-kHz checkpoints.

If you need a quick and handy crystal calibrator for your receiver, by all means refer to any number of previously published articles. However, if you have a little time and want to learn a bit more about electronics, then read on.

The circuit presented here is absolutely *not* in the category of, "wire the parts together according to the given values and layout and you will have . . ." category. Rather, adequate design data is provided so you can calculate the required component values and make a breadboard with a minimum amount of the test equipment. The end result will be a useful, working calibrator that provides 100-kHz and 1-MHz reference points. More important, you'll learn a bit about electronics and why it isn't really too formidable to design and build simple circuits rather than going to the store for another piece of gear, or slavishly copying someone else's design without the *foggiest notion* of what to do if it fails to work when it is plugged in.

the calibrator

The circuit in fig. 1, as opposed to the usual crystal-calibrator oscillator, relies primarily on the frequency stability of a low-frequency pulse generator that is used to supply impulse energy to a high-Q tank circuit tuned to 1 MHz. Although the oscillator is not crystal controlled the short-term stability will be more than adequate; long-term stability is primarily dependent upon the quality of key com-

ponents and regulation of the supply voltage.

The circuit can be broken into four easily understood sub-parts. Pnp transistor Q1 and its associated resistors (R1, R2 and R3) form an adjustable current source flowing from the collector of Q1. This current, I1, can be set to the desired value by adjusting the base voltage V1 with potentiometer R2. Current I1 will be very nearly (neglecting some fine points concerning base-emitter diode potentials):

$$I1 = \frac{E - V1}{R1}$$

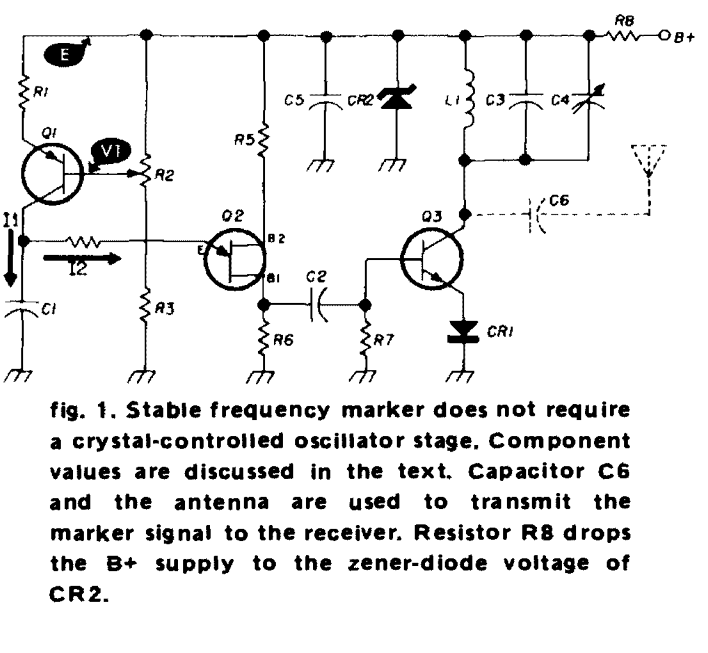


fig. 1. Stable frequency marker does not require a crystal-controlled oscillator stage. Component values are discussed in the text. Capacitor C6 and the antenna are used to transmit the marker signal to the receiver. Resistor R8 drops the B+ supply to the zener-diode voltage of CR2.

Current I1 charges C1 *linearly* to an increasing higher voltage until the threshold firing potential of the unijunction transistor is reached; at this point C1 is rapidly discharged through R4 into Q2. This process repeats indefinitely since Q2 will revert to a non-conducting state whenever the emitter voltage falls below a minimum called the *valley voltage* and current I2 is less than the *holding current.*

If an oscilloscope were used to look at the voltage across C1 it would display a very good sawtooth waveform; a linear voltage ramp due to the constant current I1 with a very fast return to near ground when Q2 conducts. Therefore, two parts of the circuit are easily understood: Q1 is a constant-current source and C1 is a timing capacitor that determines how long it takes to reach the trigger voltage of Q2.

Transistor Q2 is nothing more than a simple unijunction transistor pulse generator. Whenever the emitter voltage reaches a certain fraction of the supply voltage E, a triggering voltage is reached which causes conduction between the two base junctions, thus causing B1 to suddenly jump up to a level near the emitter voltage. If the emitter voltage drops below some specified minimum conduction between B1 and B2 ceases. Thus, looking at R6 with a scope would show a squarewave pulse every time Q2 conducts.

Transistor Q3 is normally not conducting. The base is normally at ground potential through R7, and the voltage required to make Q3 conduct is at least two diode voltage drops – CR1 and the base-emitter diode of the transistor. Therefore the base must reach at least 1.0 to 1.2 volts before Q3 will be turned on.

However, each time a pulse appears at R6 it will be coupled to Q3 base by C2 briefly turning on Q3. If C2 is fairly small it will differentiate the squarewave pulse at R6 (a voltage spike will be observed at the base of Q3).

The final portion of the circuit consists of Q3 and its resonant tank circuit. Rather than being an ordinary tuned amplifier, this circuit could be called an "impulse driven resonator." Each time Q3 conducts because of a voltage spike from C2 energy is fed to the collector tank circuit. If the tank circuit has high Q it will continue to ring at the resonant frequency even though the initial energy supply is removed (Q3 stops conducting). The initial sine-wave energy at the resonant circuit will slowly diminish in amplitude but will be replenished when Q3 fires again. If the tank circuit is tuned to 1 MHz you have a 1-MHz marker plus plenty of harmonics to use as higher band-edge markers.

circuit kinks

Although the foregoing discussion should be enough to get the circuit playing, a few more words should reduce the amount of "fiddle-time." Since you want a stable oscillator the supply voltage should be constant; a large electrolytic capacitor, C5, and an appropriate zener regulator can be used. The voltage is best chosen in the range of 12 to 28 volts dc.

Any old pnp transistor can be used for Q1. A value of 10k ohms for R2 and R3 is a reasonable starting value. Similarly, R4 isn't too critical since it is only used to limit the initial current surge into Q2; something in the neighborhood of 1000 ohms is reasonable.

Any problem likely to be encountered is in the choice of R1 and C1. A basic impulse rate that is a submultiple of 1 MHz is desirable so 100 kHz seems like a good basic frequency for the sawtooth at C1. However, current I2 *must* eventually diminish to a value less than the holding current for Q3 (on the order of 1 mA).

If I1 is greater than the holding current for Q2 (check the manufacturer's specs for this value) then Q2 will latch-up in the conducting state. Therefore the current selected for I1 must be less than the Q2 holding current or R4 made very large, which is not desirable. Similarly, to obtain a pulse rate of 100 kHz into Q2, C2 must be charged to the trigger voltage of Q2 at precisely that rate, every 10 microseconds. To pick the proper value for C2 once I1 has been selected use the equation

$$\frac{\Delta V}{\Delta T} = \frac{I1}{C1}$$

where the symbol Δ means "rate of change." More simply, the voltage change at C1 (called ΔT) is equal to the ratio of the current I1 divided by the capacitance charged, C1 in this case. Since you know I1 (or can vary it slightly) and ΔT is 10 microseconds, then the value of C1 is easily determined since the required value of ΔV is the triggering voltage for Q2. (The triggering voltage is typically 0.25 to 0.5 the supply voltage.)

The unijunction transistor Q2 is fairly non-critical as to the values selected for R5 and R6. Resistor R5 usually ranges from 300 to 600 ohms; 100 ohms is a ballpark value for R6. Since you want some differentiation of the pulse delivered to Q3, capacitor C2 might be anything from 1000 pF to as much as 0.01 μF.

Resistor R7 keeps Q3 reverse biased by draining off the reverse base-leakage current; for most small-signal transistors this leakage is so small that 100k ohms is a fair value for R7. Nearly any npn transistor will work adequately at Q3 such as an old 2N706 or 2N708.

The tank circuit (L1-C3-C4) should resonate at 1 MHz; a high-Q inductance is needed. This is easily obtained by using a ferrite toroid core and many turns of small wire. Capacitor C3 should provide most of the required capacitance, with C4 used for fine tuning. Alternatively, if L1 is variable C4 may be omitted. A wire antenna can be attached to the tank circuit to couple energy to your receiver.

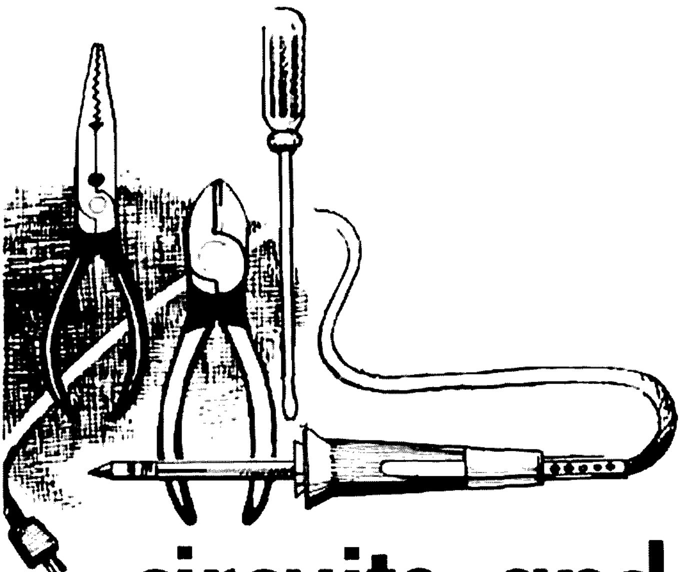
calibration

Circuit calibration may be accomplished by zero-beating the output against WWV. The primary variable will be R2 to obtain zero beat, with C4 used to obtain maximum power output. However, C4 will alter the output frequency slightly so some juggling back and forth between R2 and C4 will be needed initially.

Frequency stability will rival that of most crystal oscillators if two precautions are observed: The supply voltage must be stable, and C1 must not change capacitance radically with temperature. To a lesser extent, more expensive unijunction transistors have more stable trigger voltages that contribute a constant output frequency.

One nice feature of this circuit, aside from its low cost, is that if 100 kHz is picked as the basic pulse frequency, 100-kHz markers can be heard throughout the lower amateur bands and used as additional reference points.

ham radio



circuits and techniques

ed noll, W3FQJ

integrated circuits

Many broadcast equipment manufacturers have taken advantage of digital ICs when designing new equipment. Such devices as JK flip-flops, divide-by-12 counters, divide-by-16 counters, 4-bit binary counters, etc. have found their way into a-m and fm broadcast transmitters. Radio amateurs and amateur radio equipment manufacturers might like to take a look at some of the possibilities.

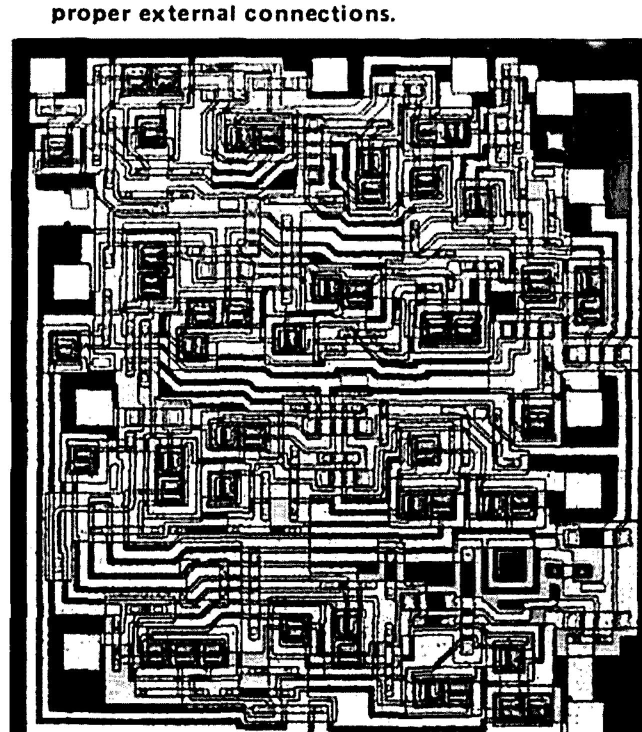
One popular a-m broadcast transmitter uses two divide-by-2 binaries, fig. 1. The crystal oscillator frequency is set in the stable frequency range from 2 to 4 MHz spectrum. If the broadcast transmitter transmits between 540 and 1080 kHz, two 2-to-1 counters provide a net count down of four. If the transmitter must operate between 1080 and 1600, a single 2-to-1 counter is used to divide the frequency.

Digital ICs of this type generate square waves but this is no problem. Output resonant circuits of appropriate Q can convert the square wave to a pure sine wave. The input to the counter should be a square wave; this can be accomplished with a simple limiter at the output of the crystal oscillator.

Some modern fm transmitters use elaborate digital counter chains such as fig. 2. This arrangement permits the frequency-modulated oscillator to operate at the fundamental frequency of the fm station. An automatic-frequency phase-lock system keeps the oscillator on the assigned frequency well within FCC tolerance (±2000 Hz on the fm broadcast band).

In one fm broadcast transmitter an output from the basic oscillator is applied

One of Motorola's newer MRTL digital ICs is the MC880P, a decade counter that also divides by five or six with the proper external connections.



to a 16,384-to-1 counter and then to a phase comparator. The crystal reference oscillator operates in the 1.5- to 2-MHz range, depending upon the assigned frequency of the fm station. This reference frequency is divided by 256 and applied to the phase comparator. The phase comparator develops a dc reference voltage which keeps the transmit oscillator on frequency.

For such a divider system, the digital ICs must be capable of operating at very high frequencies. Such units are available at low cost. For example, a 120-MHz J-K flip-flop has a unit cost of approximately $7.50. A flip-flop with a maximum input frequency of 85 MHz can be purchased for about three dollars. A divide-by-16 counter with a maximum frequency of 30 MHz can also be purchased for about

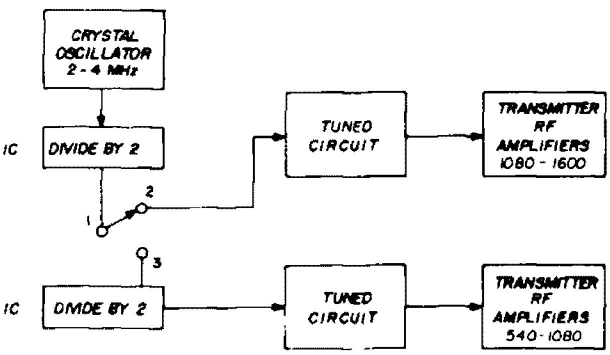


fig. 1. Circuit for using IC frequency dividers in a-m broadcast transmitters.

$7.50. A search of the surplus market may yield the same or similar units at even lower prices.

practical frequency dividers

What are some of the possibilities for amateur equipment? Amateurs are accustomed to using frequency multipliers that develop signals related harmonically to a crystal oscillator or vfo. A digital divider system permits decreases in frequency. For example, a divide-by-2 counter following a 40-meter vfo that tunes from 7 and 8 MHz will give an 80-meter output between 3.5 and 4 MHz. Two 2-to-1 dividers would provide a net division of 4 and put the signal into the 160-meter band, fig. 3.

Digital ICs that can be connected as astable multivibrators can also be made to

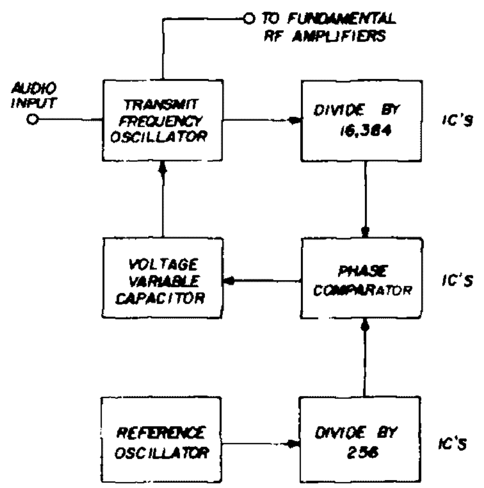


fig. 2. Integrated circuits in a phase-lock system determine the operating frequency of this fm broadcast transmitter.

multiply. Is it possible to build an IC substitute for the common frequency-multiplier chain?* The arrangement of fig. 4 might be feasible as an all-band signal source or as an accurate all-band signal source and calibrator. Two-times the 7.5-MHz master-oscillator signal provides a WWV check point at 15 MHz.

Reconstructing a sine wave from a square wave output is no great problem

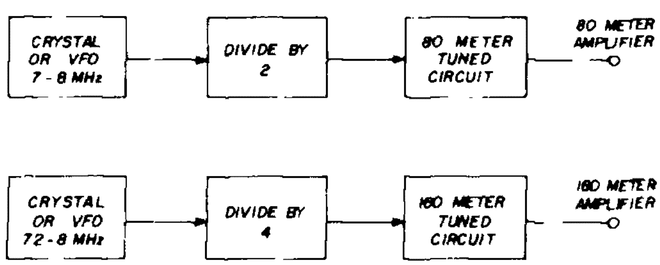


fig. 3. Two possible circuits for using IC frequency dividers in amateur radio equipment.

with resonant circuits or multisection integrators. However, you are working with pulse waveforms and their multifrequency makeup, so proper shielding and grounding is important.

This scheme is attractive for all-band QRP operation and it matches the various QRP modes of cw, a-m, dsb and phase-type ssb. There is also merit in terms of the direct-conversion receiver because you can work both down- and up-frequency from a single high-stability beat oscillator.

For vhf operation IC frequency dividers can be used to put the transmit oscillator on the transmit frequency, doing away with the multiplier chain. A possible system for 2- and 6-meter operation is shown in **fig. 5.** A divide-by-8 circuit brings a fundamental 6-meter signal down to the 7-MHz range where it is compared with a crystal oscillator, vxo or stable vfo. The two signals are applied to a phase comparator (also available as a digital IC). A dc afc voltage is developed for controlling the fundamental frequency oscillator, holding it on frequency. The fundamental frequency oscillator uses a voltage-variable capacitance diode that responds to the phase-lock afc system.

For 2-meter operation an 18-to-1 frequency divider brings the fundamental frequency down to the 8-MHz range for comparison. Otherwise, the system is the same as 6-meter arrangement. Also, this configuration is adaptable to various modes of modulation. Frequency modulation may even be accommodated. However, frequency division must be greater to prevent the modulating frequencies from affecting the phase-lock operation.

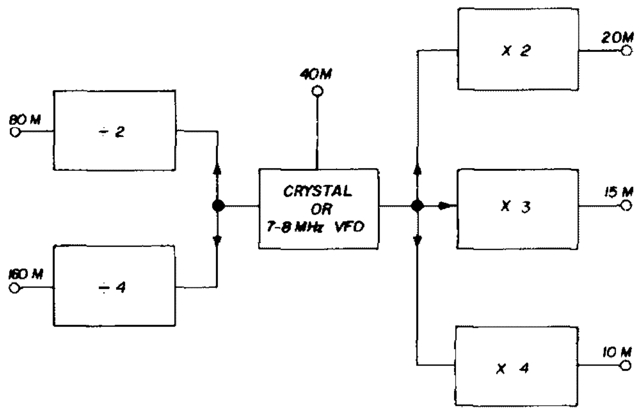


fig. 4. An all-band high-frequency signal source using a single 40-meter oscillator.

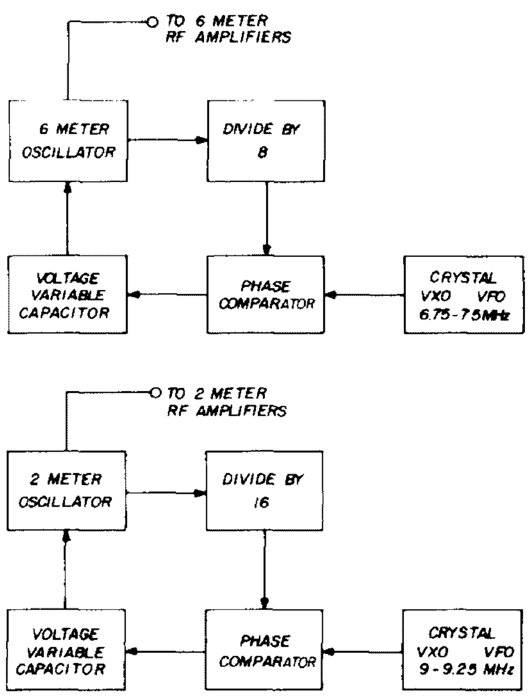


fig. 5. Using IC frequency dividers in frequency-determining systems for 6 and 2 meters.

simple rf divider

The feasibility of this frequency-control circuit was checked out with a low-cost MDTL clocked R-S flip-flop, wired externally to operate as a J-K flip-flop. The frequency divider was used with the fet circuit described in the June issue.

The output of the Pierce crystal oscillator, **fig. 6**, is fed directly to the clock input of the flip-flop. A shaping diode is connected between pin 2 and common. The flip-flop is biased with a 9-volt transistor radio battery.

A tuned fet rf amplifier follows; it is operated in class A because of limited drive from the flip-flop. Nevertheless, approximately 200 milliwatts of rf output is developed. But more important, a beautiful 80-meter sine wave may be observed at the output. A local 80-meter QRP contact using a 40-meter crystal is something a bit different from the ordinary.

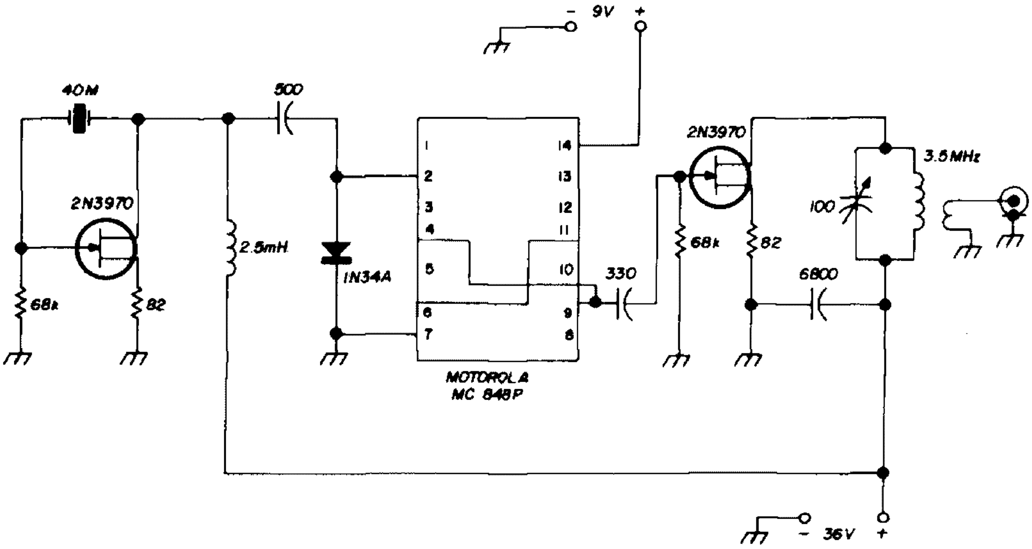


fig. 6. This QRP transmitter for 80 meters uses a 40-meter crystal and IC 2:1 frequency divider.

high-power audio

An unusual two-section high-power audio amplifier, the RCA HC-1000, is capable of delivering 100 watts into a 4-ohm load with a peak current of 7 amperes. The total supply voltage is 75 volts. An output transformer can be added to step up the impedance when required; this may be necessary if the unit is used to supply audio to a vacuum-tube modulated amplifier.

The price of the HC-1000 is in the $60 bracket. This might shake you a bit but consider what it would cost to build a vacuum-tube modulator capable of de-

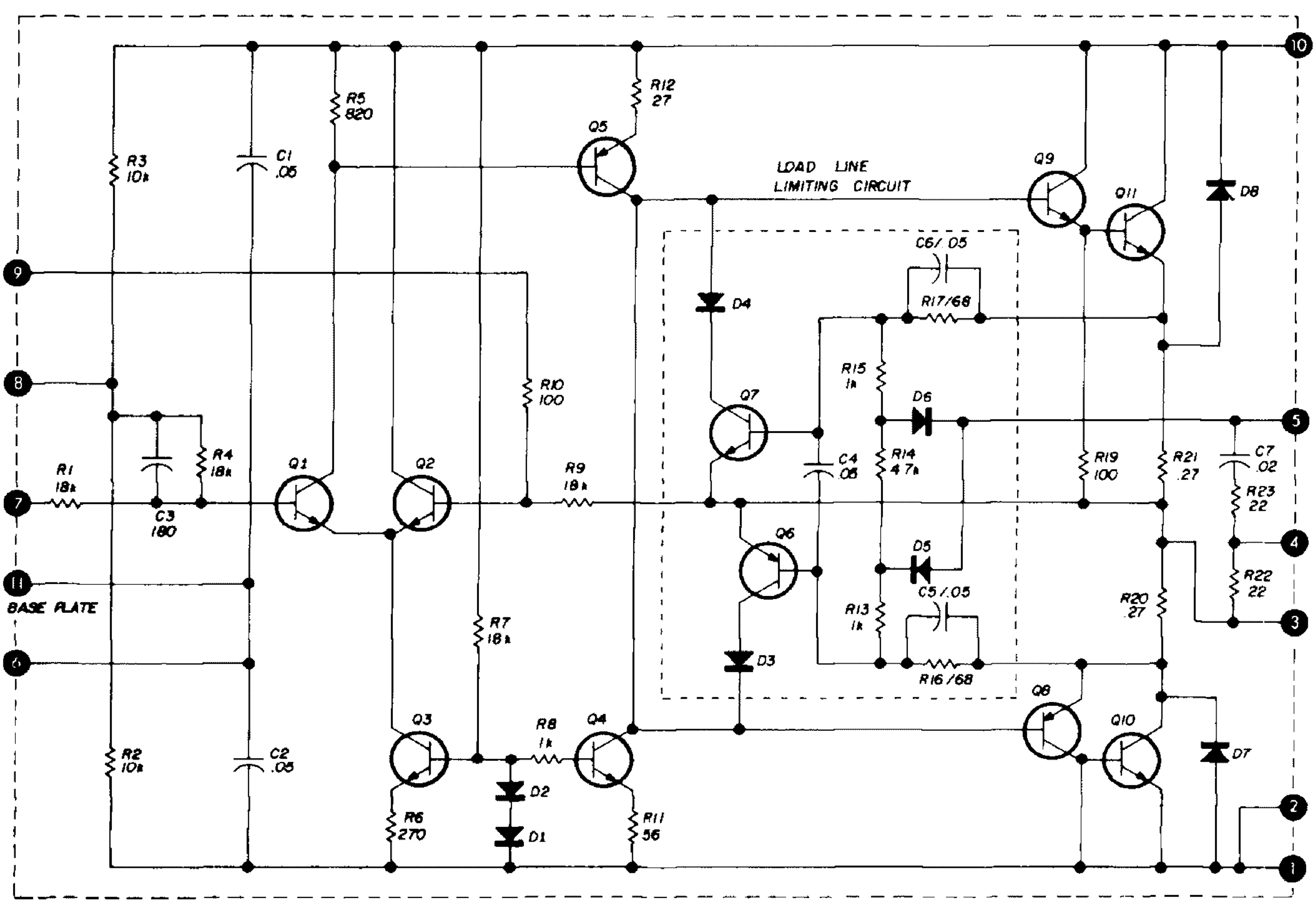


fig. 7. Internal circuit of the miniature RCA HC-1000 100-watt audio power amplifier.

livering 75 to 100 watts of audio? Consider too that the entire HC-1000 module is 3 by 3-1/2 inches and weighs but 100 grams.

The amplifier consists of two separate sections mounted on a common base plate. One section serves as the complete driver and includes 23 resistors, 7 capacitors, 6 diodes and 9 transistors. Two power output transistors and two diodes are contained in the second section.

A schematic diagram of the HC-1000 is shown in fig. 7. The input stage (Q1 and Q2) is a differential amplifier with a constant-current bias source (Q3). Note the temperature-stabilizing diodes in the base circuit of Q3.

The collector of Q1 supplies output to the base of a class-A amplifier (Q5). Transistor Q5 supplies drive to the two pairs of output transistors. The collector of Q5 is connected to the collector of Q4; Q4 is also a constant-current source but its purpose is to supply a bidirectional current in proportion to the current through Q5. It acts as a level translator to ensure that no dc offset voltage is applied to the output transistors. Such a dc voltage would be contributed by Q5 if it were not balanced out by the voltage developed by Q4.

The protection circuit within the dashed block prevents damage to the module when an improper load is placed on the output. Transistors Q9-Q11 and Q8-Q10 are Darlington pairs that develop a high-power output across a low impedance load. The input impedance to this stage is high enough to attain good gain from the class-A amplifier (Q5).

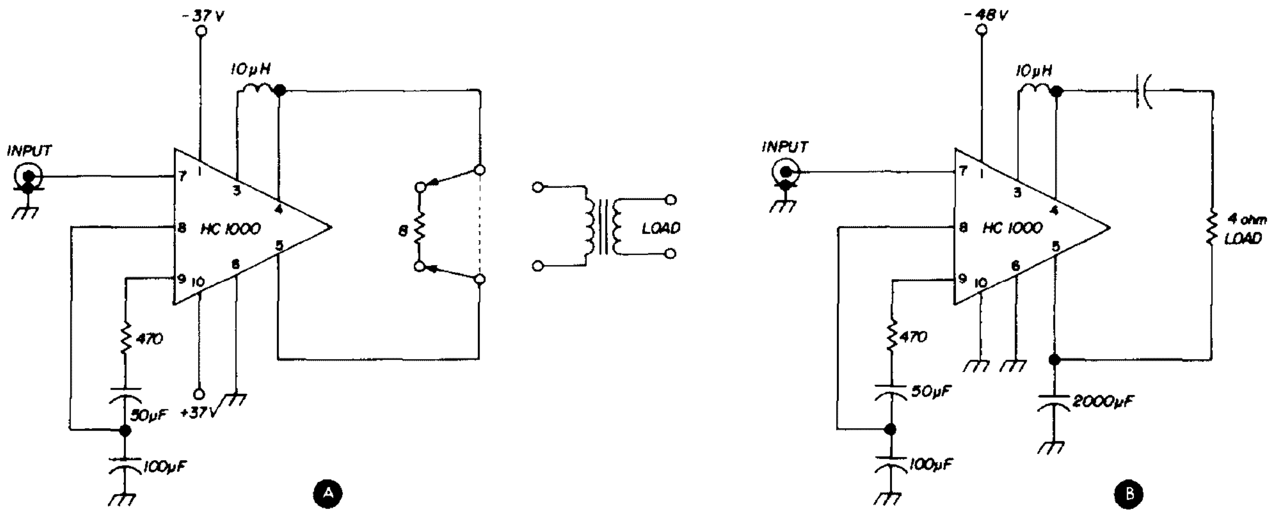
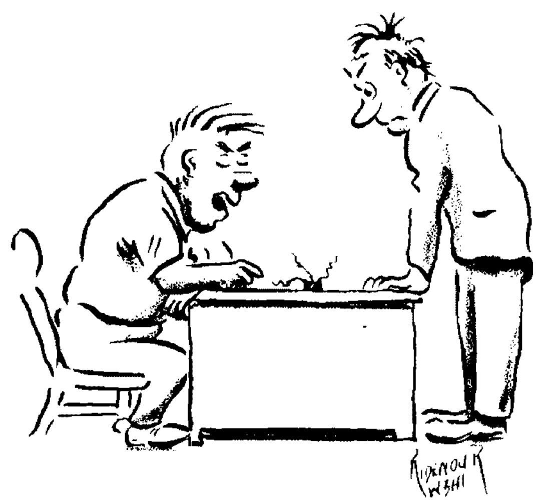


fig. 8 Practical audio power amplifiers using the HC-1000. Circuit in (A) requires both positive and negative voltages; circuit in (B) requires only negative supply voltage.

The HC-1000 circuits in fig. 8 are designed for a positive and negative power supply (fig. 8A) or a negative power supply (fig. 8B). Circuit 8A includes an output transformer for providing a proper match to a specific load resistance.

ham radio



"This used to be an integrated circuit — until my wife thought it was a spider and stepped on it!"

the ham notebook

switch-off flasher

Portable solid-state equipment that is designed for low power drain often excludes the use of "high-drain" pilot lamps. The circuit in **fig. 1**, designed by S. Thomas and described in *The Radio Constructor,* flashes the light *after* the switch is turned off. The circuit is applicable to any transistor equipment that uses a 9-volt battery. When switch S1 is *on* the capacitor is charged through the diode; when the switch is turned *off* the 6-volt lamp is connected across the capacitor, causing the bulb to flash brightly. The circuit draws negligible battery current; in Mr. Simon's unit the leakage current drawn by the capacitor was measured at 6 μA.

The capacitor cannot discharge the equipment circuits because of the diode. However, if the equipment has a large-value capacitor across the power supply, it will discharge through the diode into the lamp, increasing the length of the flash. With most equipment it would probably be possible to eliminate the diode and connect the capacitor and lamp *directly* to the negative supply lead, although this should be checked experimentally.

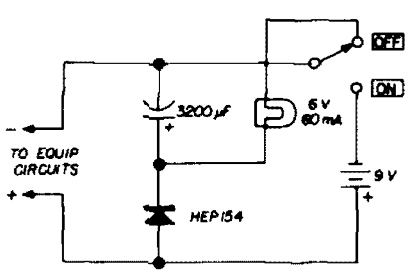


fig. 1. Simple low-drain circuit flashes when the switch is turned off. Current drain is less than 10 uA.

receiver incremental tuning for the swan 350

Although receiver incremental tuning is not an absolute necessity, at times it is a pleasant convenience. The circuit in **fig. 2** is quite basic and may be added easily to the Swan 350.

Most of the smaller component parts may be mounted on a small 6-lug terminal strip which is mounted under the left-rear mounting nut of the vfo printed-circuit board (as viewed from the front with the Swan on its back). This provides a short, direct connection between C1 and the collector of the oscillator transistor, Q1, by soldering to the lead which comes from the dial set and tuning capacitors; this lead is connected to the vfo board at the small solder blob immediately in front of the mounting screw that is used to mount the terminal strip. Use a piece of good stiff wire here to prevent any unwanted fm-ing due to vibration.

Location of the other components is not critical. In my unit CR2 and R4 were mounted on a terminal strip adjacent the voltage regulator tube, V16. The relay was mounted immediately behind the vfo box near the two elongated chassis cutouts. The *RIT* pot was mounted on the front panel between the dial window and the plate tuning control. Although I

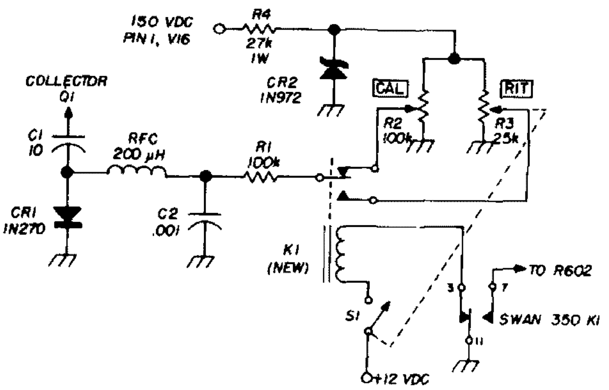


fig. 2. Incremental tuning circuit for the Swan 350. Relay K1 is a miniature 12-volt spdt relay such as the Potter & Brumfield KM5D. CR2 is a 30-volt, 400-mW zener.

mounted the *calibrate* pot on the front panel above the RIT pot, it may be mounted internally on a small bracket. The wires were routed through the small chassis hole near the vfo amplifier tube.

Calibration is accomplished by first setting both pots to the middle of their ranges. Feed in a crystal-calibrator signal and pull out on the RIT knob, leaving it at center. Now zero beat the calibrate signal with the main tuning knob. Push the RIT knob in (*off*) and adjust the *calibrate* pot for zero beat. This insures the transmit and receive frequencies are the same with the RIT function disabled while allowing manual control of the *receive* frequency with the RIT on.

The installation of the circuitry upsets the dial calibration very slightly; it was found the dial-set capacitor easily restored calibration. With the circuit constants shown, ± 3.5 kHz frequency deviation was obtained. With this amount of shift you may take a station from zero-beat and move the signal out of the passband of the filter in either direction.

Paul Pagel, K1KXA

crystal-controlled multivibrator

The crystal-controlled oscillator circuit shown in fig. 3 uses no tuned circuits. Since there is no reactance in the circuit except for the small gimmick capacitor, the circuit will oscillate at any frequency from 2.5 kHz to 15 MHz by simply changing the crystal. Rf output voltage is approximately the same as the supply voltage, which can be from 6 to 25 volts.

I have used this circuit as a wideband frequency spotter, as a marker generator (output is rich in harmonics), and as a crystal activity and stability tester.

When building the oscillator, keep all interconnecting wires as short as possible. Although it is not required in every case, you may need the gimmick capacitor to obtain oscillation. The gimmick is made by twisting together two 2-inch pieces of insulated hookup wire; make sure the two wires don't short together.

Mike Centore, WN2MQY

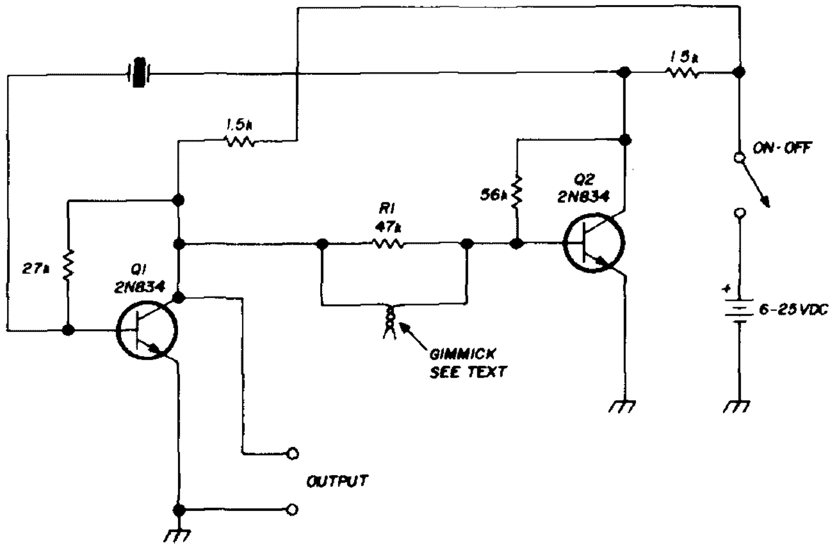


fig. 3. Crystal-controlled multivibrator uses no tuned circuits but will oscillate at any frequency between 2.5 kHz and 5 MHz by simply changing the crystal.

versatile resistor decades

Accurate decade standards may be assembled at considerably less than their usual cost by using one of the switching arrangements shown in **fig. 4** and **5**. The precision components represent the greatest cost of a decade standard; therefore, any reduction in the number of components directly reduces cost.

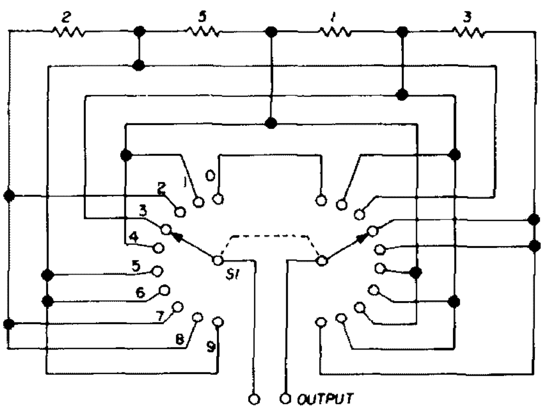


fig. 4. Zero through 9 ohms are obtained by switching various combinations of four standard resistors. For higher resistance values the resistors are increased by factors of ten. The switch is a 2-pole 10-position rotary.

Only four precision resistors or capacitors are required in this design, instead of the usual ten. This savings may be used to simply reduce cost, or if you wish, to buy closer tolerance components for a more accurate decade.

Several construction precautions should be observed. Low-resistance decades require very low contact resistance switches to preserve accuracy. This requirement may be met by parallel wiring two decks of a conventional rotary switch, instead of single decks shown in **fig. 4**.

In capacitance decade's low-value capacitors must be well spaced from all metallic components to reduce errors due to stray capacitance effects. Also, wide spacing between capacitor terminals and jumpers is required. In the low-capacitance and high-resistance decades, rotary switches with ceramic insulation are recommended.

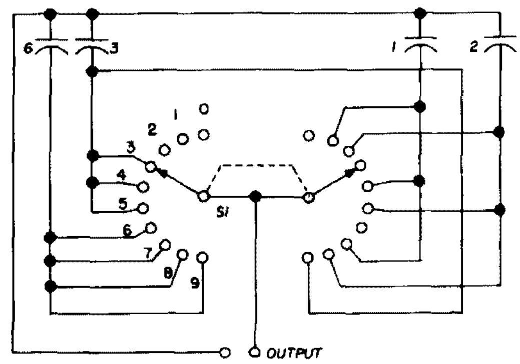


fig. 5. Nine individual capacitance values are provided by switching combinations of four capacitors. Switch is a 2 pole 10-position rotary.

The economical decade switching systems shown here may be extended to the design of other useful decades including inductance, zener diodes and back-to-back diode clippers.

Gene Brizendine, W4ATE

vfo buffer amplifier

The solid-state vfo I described in the August, 1970 issue of *ham radio* was designed to provide a high-quality stable signal at low output to replace crystals in low-power solid-state transmitters. It is essential in any vfo that oscillator loading be as small and constant as possible (to produce a chirpless signal). This was achieved in my vfo by voltage regulation and by using a low-load buffer stage. The third stage was designed to match the high-impedance output of the jfet buffer stage to the low-impedance input of a solid-state transmitter. Consequently, there is little amplification of the oscil-

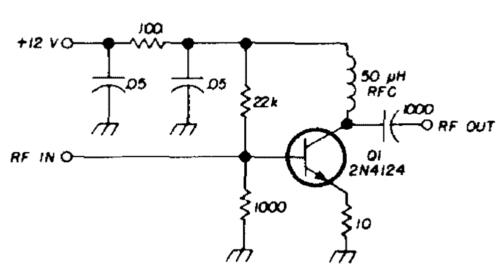


fig. 6. Solid-state vfo amplifier provides constant high-impedance load to the vfo. RFC1 is a Millen 34300-50.

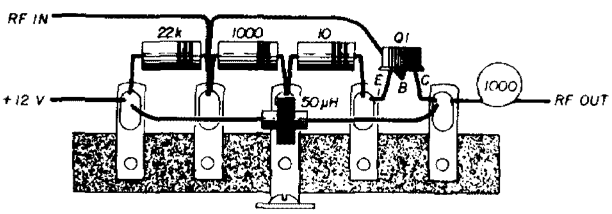


fig. 7. Vfo buffer amplifier can be built completely on a 5-lug terminal strip.

lator signal; output voltage is about 1.5 volts maximum.

If you need more output for this or similar solid-state vfo you may be interested in adding the simple one-stage amplifier shown in **fig. 6.** This amplifier will provide plenty of power to the driver stage of a transistor transmitter up to the five-watt class, and can be built for under $3.00 using all new parts. This circuit will fit nicely into the W3QBO vfo minibox; or, it can be built into the associated transmitter.

The components can all be neatly mounted on a 5-lug terminal strip as illustrated in **fig. 7.** The schematic shows a 100-ohm resistor and two .05-μF bypass capacitors for power-lead decoupling, a practice that should always be followed. The Motorola 2N4124 transistor sells for 60 cents; many other transistors (such as the Motorola 2N3904) will work well in this circuit but usually they will cost much more.

C. Edward Galbreath, W3QBO

vhf a-m modulation monitor

If you have a dc-coupled oscilloscope in the shack, such as the EICO 460, here is a low-cost and simple method of keeping *track of your percentage* of modulation. At vhf the usual techniques of envelope patterns and trapezoidal patterns become very difficult, if not impossible. The method shown in **fig. 8** is based on the fact that the plate voltage applied to an amplitude-modulated stage varies between zero and twice the dc supply voltage for 100% modulation of a *properly adjusted* a-m transmitter.

Connect the vertical scope input to the modulated dc which feeds the modulated rf stage. A resistive voltage divider (no capacitors) is necessary if the voltage is high enough to be dangerous. Select a convenient sweep rate of about 100 Hz or so. Position the horizontal trace so it coincides with one of the lower graticule lines. Apply plate voltage to the modulated stage (with no modulation) and adjust the vertical gain control to

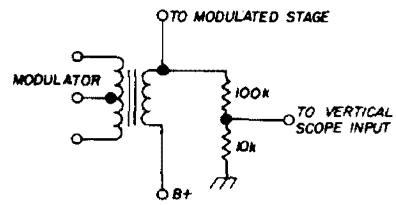


fig. 8. Resistive divider is used to pick off signal for a-m modulation monitor.

move the horizontal trace to the center graticule line. The scope is now set up to show modulation amplitudes varying around the dc supply voltage.

Audio modulation gain should be adjusted to keep the scope pattern between the limits of the lower (zero volts dc) graticule line and the corresponding graticule line above the center. **Fig. 9** shows the relationships. For my equipment, with a 750-volt dc power supply, I chose a resistive voltage divider of 100k and 10k. A suitable ratio permits convenient measurement of the final dc supply voltage without exposure to lethal high voltages.

Harry Ferguson, K7UNL

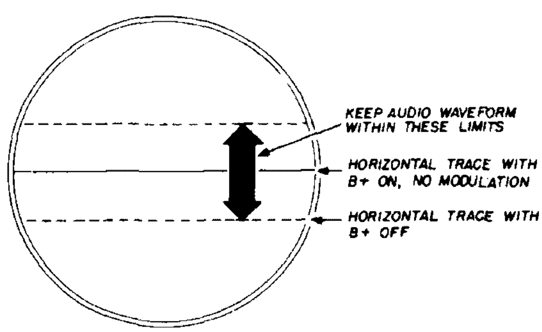
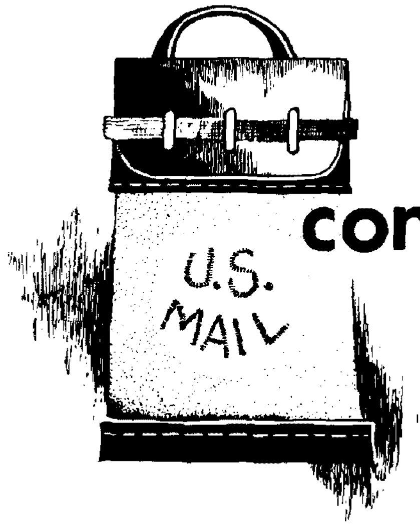


fig. 9. Modulation percentage is less than 100% if waveform is between limits shown here. Equipment setup is described in text.



comments

QRP technique

Dear HR:

W6NIF's article on QRP technique in the December issue was interesting and valuable. But it seems to me that QRP has another facet which might be of immediate concern to more amateurs. Personally, I have the utmost respect for those, like Al Wilson, who persist with real flea power. I wish I had the time and patience to do so. However, to me *practical* QRP means power inputs of the order of five watts.

Now that the southern gentlemen at Ten-tec have introduced so many of us to this fascinating QRP world we may learn what we should have known for years. Most of us regularly use from five to one-hundred times more power than we need, or ought to use. We QRM the world for our ego's sake.

For almost a year now, I have not used *more than* eight watts dc power input to the final stage, and I have had more fun in the game than ever. Using both a Ten-tec rig, and a number of homebrew jobs, I have enjoyed as many *solid* QSOs as I formerly did with a ninety-watter. In fact, most folks worked do not know I'm QRP until I tell them. And I use a most unpretenious antenna, an 80-meter zepp, up only about 25 feet. Folks with *good* antennas, and perhaps more skill, do better. But I've worked 20 countries, 24 states, and all parts of Canada, and relished it immensely. Furthermore, I don't feel like an "ugly American."

To me, this is *amateur* radio, as contrasted with the expensive, tinseled world of "commercial" ham radio that one hears, particularly in the phone bands, where power seems valued for its prestige alone. Less-than-one-watt QRP is interesting and may constitute a valid achievement. But few of us have the time, the patience and perhaps the skill for this. All of us, however, could cut our power input to one-tenth of its present level, have as many fine QSOs as ever and thereby vastly improve the American amateur's electronic world image. This in itself, in these times, would seem worthwhile.

C. F. Rockey, W9SCH
Deerfield, Illinois

fire extinguishers

Dear HR:

The January article, "Fire Protection in the Ham Shack," is a very important and timely subject. However, I question the statement about Freon not being as toxic as carbon tetrachloride is. My experience indicates that Freon is highly toxic when exposed to high temperatures.

For a number of years I was in the air-conditioning trade. In one case a Freon-refrigerant discharge line broke in an air-conditioning system in the basement of a hotel; in the same room with

the broken line there were several gas-fired hot-water heaters. When the freon broke down in the flame it formed phosgene gas — the same results produced by carbon tetrachloride. I was extremely sick from being exposed to the fumes, and it was some time before I was well again. I would suggest ventilation as soon as possible if using a Freon-type fire extinguisher; in fact, in my book, this goes for the use of any aerosol cans.

Orville Gulseth, W5PGG
Clarksdale, Mississippi

frequency-sensitive resistors

Dear HR:

I am a sales engineer, and the following comments come from 18 years of selling resistors. Contrary to popular belief, deposited carbon-film, metal-film, and the like, are *not* non-inductive resistors. After the resistive material is deposited on the ceramic core the resistive material is cut into a spiral, with 12 or more turns before the desired resistance value is obtained. The result is a 12-turn coil of resistive material plus some capacitance thrown in for good measure.

If you want a true non-inductive resistor you must tell the manufacturer what frequency you will use, and then stay close to it. For all-around general-purpose non-inductive resistors for all frequencies the old carbon-composition resistor is hard to beat.

Ed Aymond, W5UHV
Dallas, Texas

more uses for the ST-6 RTTY demodulator

Dear HR:

Although the ST-6 was not designed with computer work in mind, the time constants are adequate even for 220 Baud. Computers use 8-level 100-speed teleprinters such as the model 33 or 37. This is 110 Baud, considerably shorter pulses than the normal 5-level 100-speed teleprinters used on many MARS frequencies.

The ST-6 operates in a normal manner even at 200 Baud, with the change of two resistors and one capacitor in the low-pass filter. I am using such a modified unit at home in conjunction with a cathoderay type computer terminal.

The ST-6 can also be used with the AK-1 AFSK unit (*QST,* February, 1965) and a phone patch as a computer interface between the phone line and the computer terminal. (The AK-1 also performs in excess of 220 Baud.) Such use would not appeal to many hams at this time, but the possibility is always there if you need it.

Irv Hoff, W6FFC
Los Altos, California

motrac receivers

Dear HR:

In the December, 1970 issue, the *ham notebook* recommended the Motorola Motrac receiver for use in repeaters. It should be noted that there are two basic types of Motrac receivers.

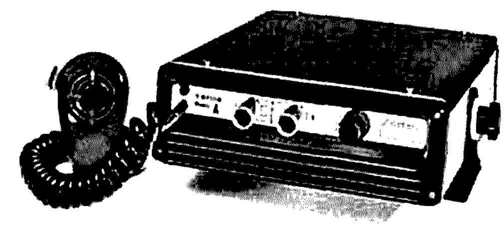
The type-L receiver uses a bipolar rf amplifier stage and is a poor choice for just about any fixed-station application. Its desensing characteristics are 5 to 10 dB inferior to the tube-type G receiver, and 10 to 15 dB worse than the old type-A receiver. The intermod characteristics are even worse: 10 to 15 dB more than the G, and 20 to 25 dB more than the A. The later type-M receivers use a fet mixer in the front end; it is about equal to the old type-A in desensitization, and far, far better in the intermod department.

The order of preference for Motorola receivers is the type-M (new Motrac), type-A (wide-chassis tube type), type-G (narrow-chassis tube type), and last, the type-L (early Motrac).

J. A. Murphy, K5ZBA
Tulsa, Oklahoma

new products

two-meter fm rig



Simpson Electronics has announced a new low-cost vhf-fm mobile rig for the amateur who wants to operate on two-meter fm. The 6-watt *Simpson Model A* features four channels, sensitivity of 0.6 μV for 20-dB quieting and separate plug-in transmitter and receiver boards with easy access to all tuning and metering points. The solid-state design uses integrated circuits and diode frequency switching.

The squelch is fully regulated and compensated with squelch sensitivity at 0.25 μV or less. Intermodulation interference rejection is 70 dB minimum. Spurious response attenuation is greater than 60 dB. Selectivity, furnished by ceramic filter, is 13 kHz at 6 dB down and 36 kHz at 60dB down. Audio output is 2 watts with less than 10% distortion. Current drain on receive: squelched, 250 mA; unsquelched, 480 mA.

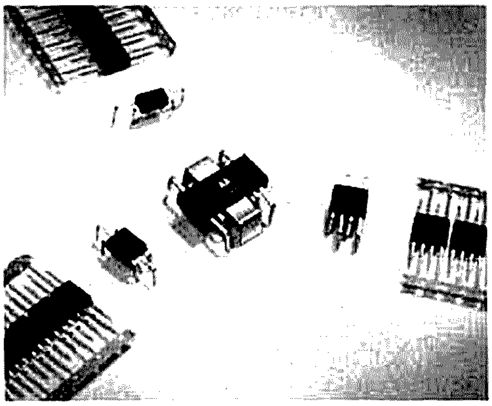
The transmitter has ±5 kHz deviation for 100% modulation at 1 kHz and includes instantaneous automatic deviation limiting. Transmitter audio response is +1 dB to -3 dB from 300 to 3000 Hz. Tuning range is 144 to 148 MHz.

The Simpson *Model A* comes complete with mounting cradle and press-to-talk microphone with coil cord. Simpson furnishes crystals for operation on 146.34/146.94 and 146.94/146.94 MHz. The Simpson *Model A* is priced at $249, including four crystals. For more information write to Simpson Electronics Inc., 2295 Northwest 14th Street, Miami, Florida 33125, or use *check-off* on page 94.

audio sweep generator

The new Rameco ASG-1 audio sweep/signal generator was developed for use in the frequency range of zero to 100 kHz. Its prime purpose is to display the response characteristics of either active or passive circuits on a standard oscilloscope. Both swept and cw modes of operation are provided; sweep width is variable from a few Hertz to 100 kHz in a single sweep. Output is adjustable from 0 to 5 volts peak to peak. A synchronized ramp output with adjustable amplitude is supplied for driving the horizontal input to the oscilloscope. If triggered operation is desired, generator blanking pulses are available. Sweep time is variable from 20 milliseconds to 20 seconds. $195.00 from Rameco Corporation, Post Office Box 580, Deerfield Beach, Florida 33441. For more information, use *check-off* on page 94.

motorola functional circuits



Motorola's recently introduced line of low-cost *functional circuits* now totals twenty devices. Included are flip-flops, voltage regulators, general purpose and audio amplifiers and an electronic attenuator.

Functional circuits provide common circuit functions where discrete transistors cannot economically be used. These are generally circuits where matching or balance or component density can better be achieved on a single chip, and include complex digital and linear functions such as multipliers and demodulators.

Functional Circuits are supplied in new, low-profile, plastic packages specially designed to use minimum printed circuit board area yet maximize pin spacing for easy, reliable soldering. The twenty functional circuits are listed below:

For further information, write to Technical Information Center, Motorola Semiconductor Products, Inc., Post Office Box 20924, Phoenix, Arizona 85036, or use *check-off* on page 94.

application	function	device number	price (1-9)
audio outputs	.25 W out audio	MFC4000A	$1.35
	1.0 W out audio w/preamp	MFC6070	$1.55
	1.0 W out audio w/preamp	MFC8010	$2.94
	4.0 W out audio w/preamp	MFC9000	$4.57
	2.0 W out audio w/preamp	MFC9010	$3.53
	2.0 W out audio w/preamp	MFC9020	$3.25
audio drivers	class-A driver	MFC4050	$1.26
	class-B driver	MFC8020	$2.38
general purpose amplifiers	wideband gain block	MFC4010A	$1.26
	fm i-f differential amplifier	MFC6010	$1.08
	flexible differential amplifier	MFC8030	$1.65
	low noise preamp	MFC8040	$2.13
digital	single-toggle flip-flop	MFC4040	$1.35
	double-toggle flip-flop	MFC6020	$2.70
	double-toggle flip-flop w/&reset	MFC6050	$1.65
	J-K flip-flop	MFC8050	$1.98
voltage regulators	8-30 V regulator 1 W, 200 mA	MFC4060	$1.41
	8-30 V regulator 1 W, 200 mA	MFC6030	$2.77
	8-30 V regulator 2.5 W, 200 mA w/current regulator	MFC9030	$3.60
control circuit	electronic attenuator	MFC6040	$1.98

vhf fm catalog

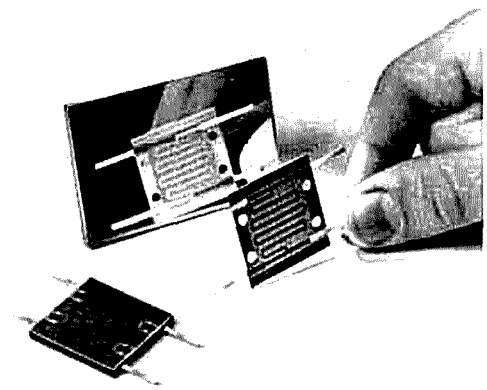
The recently published 1971 catalog from Gregory Electronics features a complete listing of equipment for the vhf fm operator including General Electric Progress, Pre-Progress and Pacers; Motorola Mobile units, Motrac, Motorcycle Dispatcher, etc; as well as commercial equipment manufactured by Dumont, RCA and Bendix. Gregory Electronics has a wide stock of fm equipment in stock, including gear for low-band (30 to 50 MHz) mobile and base stations, high-band (150-174 MHz) mobile and base stations and uhf (450 to 470 MHz) mobile and base stations, plus remote controls and various parts and accessories. For the serious vhf fm operator the new 1971 Gregory Catalog represents one of the most complete listings of vhf fm equipment, with descriptions, available to the amateur. For your copy, write to Gregory Electronics Corporation, 249 Route 46, Saddle Brook, New Jersey 07662, or use *check-off* on page 94.

motrac tone kit

The addition of sub-audible continuous-tone squelch to Motorola Motrac radios has been made easy by the new Alpha MK-21 Motrac tone mounting kit. The MK-21 makes possible rapid conversion of previously non-toned Motrac radio units, a heretofore difficult problem. Radio systems employing so called "NON PL" Motracs can be updated and made more efficient, relieving the user from listening to co-channel congestion. Because the Alpha tone unit is completely compatible with existing "Private Line," "Channel Guard," "Quiet Channel" and other sub-audible tone devices, the non-tone Motrac can be put into service in existing toned systems, thus extending the useful life of these fine radios. Also available from Alpha is an optional two-frequency tone unit that allows for selective control of repeaters, base stations, etc. The MK-21 uses the plug in the ST85H encoder or the SS80H encoder/decoder. Simple step-by-step installation instructions are provided, and a wide range of EIA tone frequencies is available.

For information write to Alpha Electronic Services Inc., 8431 Monroe Avenue, Stanton, California 90680 or use *check-off* on page 94.

uhf quadrature coupler



Multiple-device combining, transmission continuity and mismatch isolation are three major performance advantages afforded by the Motorola MIC5830-31, 3-dB quadrature coupler.

Mismatching of transmitter ports, for example, is reported no longer a problem because the application of a reflected signal at either of the output ports results in signals at the input port attenuated by 20 dB. Insertion loss is as low as 0.25 dB maximum, affording greater output power; phase balance is ±1.5° maximum, furnishing smaller combining losses and less distortion.

Efficiency in higher frequency designs, such as broadband military and ECM equipment is also heightened through elimination of power-robbing, load-balancing passive components which constrict bandwidth. Usable frequency capability ranges from 225 to 400 MHz and 450 to 512 MHz.

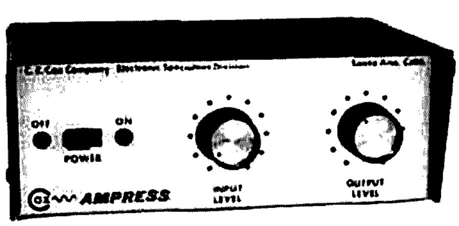
The stripline, broadside devices are constructed from Teflon fiberglass board and sealed with a low loss, low dielectric compound. Small size is achieved by meandering the coupled lines.

For complete specifications write to Motorola Semiconductor Products, Inc., Box 20912, Phoenix, Arizona 85036, or use *check-off* on page 94.

hep functional circuits

The new series of Motorola HEP audio functional integrated circuits is designed to provide a complete audio system for experimenter projects. The HEP functional circuits are available in three power output levels: HEP type C6004 provides 1 watt of audio and is priced at $2.60; type C6005 provides a minimum of 2 watts and is priced at $4.35; the 4-watt HEP C6006 sells for $5.60. All these units are high-gain devices, include a preamplifier, and will provide full-rated output with an input signal in the 10-to 15-mB range. The circuit can be used with 4-, 8- or 16-ohm speakers. A mail-order source for Motorola HEP devices is Circuit Specialists Company, Box 3047, Scottsdale, Arizona 85257. For more information, use *check-off* on page 94.

ampress speech processor



The C. E. Cox Company has introduced its new solid-state *Ampress* speech processor, a high-performance low-cost unit designed to increase talk power and improve speech quality. The *Ampress* is not a speech clipper, but a variable-gain preamplifier that uses controlled negative-feedback to produce a constant output level automatically. The circuit operates with very fast attack time to accomodate transients. In addition, the *Ampress* incorporates a "speech enhancement" feature which amplifies voice signals selectively, with frequency response controlled by the compression function. The result is a crisp penetrating signal with carefully exaggerated speech harmonics and high-frequency components to punch through heavy QRM.

The *Ampress* will work with all amateur rigs, a-m, fm and ssb, and comes in two attractive models: the CCA-1 which operates on three pen-light cells, or the CCA-1R which operates on three rechargeable cells and has an internal current regulator and external ac adapter/charger. The price of the CCA-1 is $29.95; the CCA-1R is $39.95. For more information, write to the C. E. Cox Company, 2415 Broadway, Santa Ana, California 92707, or use *check-off* on page 94.

understanding solid-state circuits

This new book by Norman Crowhurst presents all the information one might need to become acquainted with practical solid-state circuits. The author has cut away all the unnecessary verbiage on semiconductor theory and gotten right down to the circuits themselves. First there's a brief discussion of the various types of semiconductors and their operating modes; then comes a thorough but practical discussion of linear amplification. The author covers dc and ac feedback, current gain, controlling current and voltage gain, phase inversion and deals with typical everyday circuits.

The next chapter covers power amplification using single and push-pull stages, complementary symmetry and protection circuits. Then comes a thorough treatment of feedback and its applications, fet stabilization, high-frequency losses, equalization, rolloff and filter alignment. Sinusoidal oscillators and function generators are also covered in detail. Other subjects included in this helpful book are gain-controlled amplification, limiting and compression, expansion, attenuators, logic and integrated circuits. 192 pages. Over 150 illustrations. $7.95 ($4.95 paperbound) from Tab Books, Blue Ridge Summit, Pennsylvania 17214, or use *check-off* on page 94.

two-meter transceiver

The new solid-state *Comcraft CTR-144* two-meter transceiver combines all the most desirable features possible into a small portable unit. The double-conversion receiver has crystal-controlled first conversion for maximum stability; mosfets are used in the rf amplifier and mixer stages. Noise figure of the receiver is 4 dB. A three-pole Butterworth filter in the front end provides more than 80-dB image rejection, eliminating interference from the aircraft band. An automatic noise limiter and adjustable squelch are built in. I-f bandwidth is 18 kHz.

The transmitter frequency may be controlled by crystal or by the built-in vfo; the crystal oscillator uses standard 8-MHz crystals. The *CTR-144* vfo operates at 7 to 9 MHz; vfo output is mixed with 65 MHz to produce 72 MHz output – this is doubled to 144 MHz. Receiver and transmitter vfo tuning are completely separate.

The only transmitter adjustments needed are output tuning and load; a front-panel switch provides selection of either a-m or fm. Power input is 12 watts; power output, 6 watts. The *CTR-144* may be operated on its internal 117 Vac power supply, or any 12- to 18-volt dc source capable of 2 amps (the receiver requires only 100 mA). An optional clip-on rechargeable battery pack permits portable operation.

The baked-on epoxy finish of the Comcraft *CTR-144* resists scratching and other marring. The panel is brushed aluminum with lettering permanently anodized in. The circuit boards are military-style glass-epoxy types with solder-plated terminals. The *CTR-144* is priced at $389.95, including a crystal for 146.94 MHz. For more information, write to Comcraft Company, Post Office Box 266, Goleta, California 93017, or use *check-off* on page 94.

foreign language QSOs

The use of foreign languages by American amateurs is increasing rapidly. The primary motivation is probably the pleasure afforded by exchanging greetings with a foreign operator in his native language. Europeans have always used amateur radio as an effective means for practicing English, and, now that ssb has made voice contacts so easy, language partners are available who wish to learn English and are willing to trade German, Spanish, French, etc., for practice in English. For the ham with language training or interest, *Foreign Language QSOs* add a new dimension and a satisfying proficiency to his hobby.

The *Foreign Language QSO* tapes have several advantages: genuine ham jargon is used throughout; the tapes are prepared by native hams who know the right phrases actually used.

The Armed Forces method of language teaching is imitated; you advance through dialogue and questions and answers, thereby increasing your vocabulary and grammatical ability. Also, sufficient material is provided to challenge the person who already has studied a foreign language and wants to sharpen his abilities, especially for amateur radio use. In addition, a complete English translation of all text, and basic information about the structure of the target language is provided.

Foreign Language QSOs are available in tapes or cassettes, English-Spanish or English-German. For more information, write to Foreign Language QSOs, Post Office Box 53, Acton, Massachusetts 01720, or use *check-off* on page 94.

More Details? CHECK–OFF Page 94

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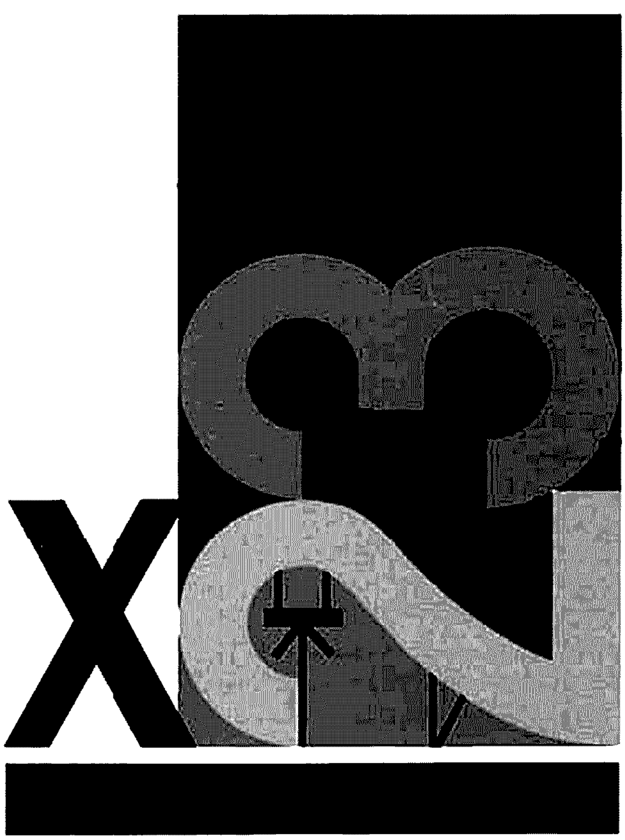
AUGUST 1971

this month



frequency
multipliers

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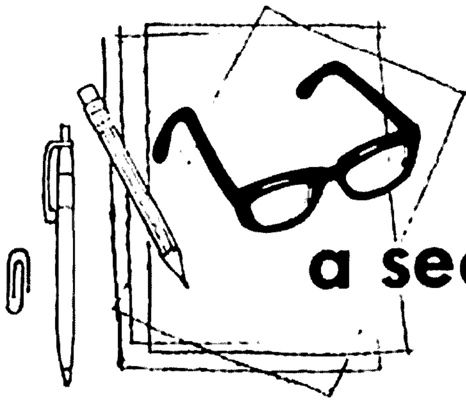


a second look

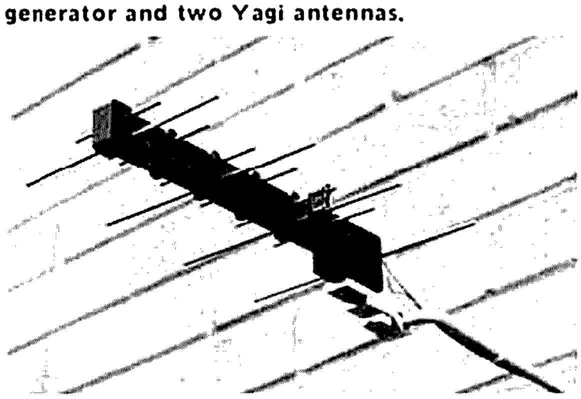
by jim fisk

The latest application for radio is a new system developed by Sangamo Electric for automatically reading electric, gas and water meters from a cruising utility truck. The new system combines the functions of meter reading, communicating the data, and preprocessing the information on magnetic tape. The tape is used to prepare monthly bills.

The utility meters used in the automatic system are conventional watthour, gas and water meters which are equipped with an spdt switch on the numerical registers; such meters are commercially available today. The electrical pulse from the switch is accumulated in a digital counter; the output from the counter is fed into a transponder mounted on the side of the building.

The transponder is the heart of the system. It is a completely passive unit that radiates no rf energy except when interrogated once a month by a cruising utility truck. The transponder is activated by a 915-MHz signal from the truck. As long as the received signal is above a preset threshold, it re-radiates the second harmonic back to the truck's highly-directive receiving antenna.

Data transponder consists of harmonic generator and two Yagi antennas.



The 1830-MHz signal from the transponder is keyed on and off by the 4-kHz pulse train from the data accumulator. A complete message from up to 12 utility meters can be received within less than 30 feet when the truck is traveling at 20 mph.

The transponder consists of two Yagi antennas on a single boom as shown in the photograph. The 4-element 915-MHz antenna is mounted below the 6-element 1830-MHz unit. The radiators of the two antennas are connected through a harmonic generator; this is the small unit in the photograph located between the second elements of each antenna.

Pilot tests being conducted in Springfield, Illinois with a prototype system indicate that, on a one-shift basis, a cruising utility truck can automatically record data from 100,000 meters per month. This represents approximately one-tenth the cost of present meter-reading methods.

sweepstakes

The prize drawings for the *ham radio* sweepstakes are taking place as this is being written. The grand prizes, a Signal/One transceiver and Delta Seventy linear amplifier, were won by W2ANB in Slingerlands, New York. W2CNB, W4YPC and WB8DUO are the proud new owners of Varitronics IC-2F fm transceivers. Fifty other winners will receive the famous RSGB "Radio Communications Handbook." We will have complete details in the September issue.

Jim Fisk, W1DTY
editor

frequency multipliers

A complete discussion of frequency multiplier circuits, including vacuum tubes, bipolar transistors, field-effect transistors, diodes and integrated circuits

Hank Olson, W6GXN, Post Office Box 337, Menlo Park, California 94025

Radio amateurs have used frequency multipliers in receivers and transmitters for many years. The octave relationship of the originally designated amateur high-frequency and vhf bands (1.75, 3.5, 7.0, 14.0, 28.0, 56.0, 112.0, and 224 MHz) made doublers very popular in the era before World War II. Also, since the time when crystal control was first introduced frequency multipliers have been needed because of the maximum frequency limit of piezoelectric crystals at any given state of the art. In the 1930s crystals were limited to fundamental-mode types; crystals above 15 MHz were generally not available to radio amateurs. Therefore, amateurs who wanted to use crystal control on 28 MHz used a doubler from 14 MHz or two doublers from 7 MHz.

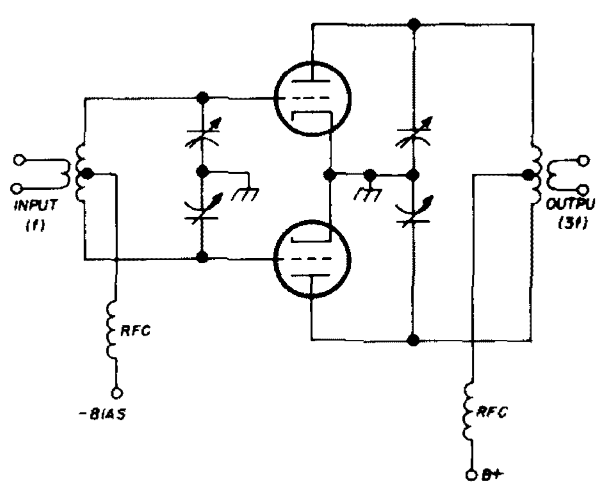


fig. 1. Vacuum-tube push-pull triple curcuit.

table 1. Vacuum-tube conduction angle for different frequency multiplication ratios.

harmonic number	conduction angle
2	90-120°
3	80-120°
4	70-90°
5	60-72°

Today, 7th- or 9th-order overtone crystals are available for frequencies up to about 200 MHz. This makes frequency multiplication less imperative. However, variable crystal oscillators (vxo) work best with fundamental mode AT-cut crystals (generally available only up to about 22 MHz) so we will still find frequency multipliers in use. Also, multiplying up from a frequency standard, generally in the 1 to 5 MHz region, is often required to make calibrations at higher frequencies.

Until 20 years ago vacuum tubes and point-contact diodes were the only devices available as frequency-multipliers. For this application the vacuum tube was usually operated in class C with the cutoff bias set so the conduction angle was considerably less than 180°. This type of frequency multiplier is covered in detail by Terman.[1] The higher the multiplication factor, the narrower the conduction angle must be for best multiplication efficiency; this is shown in **table 1.**

Two variations of the vacuum-tube multiplier are available to improve performance; both use balanced tubes. The push-pull multiplier, **fig. 1,** cancels out even harmonics so is useful as a 3-, 5- or 7-times multiplier. This circuit looks just like a push-pull class-C amplifier except that the cross-neutralization capacitors are removed since the output is on a different frequency than the input. The push-push multiplier, **fig. 2,** cancels the fundamental and odd harmonics so is useful as a doubler, quadrupler, 6-times multiplier, etc. Anyone who worked the vhf bands in the 1950s will remember the multi-stage push-push and push-pull 6J6 multipliers.

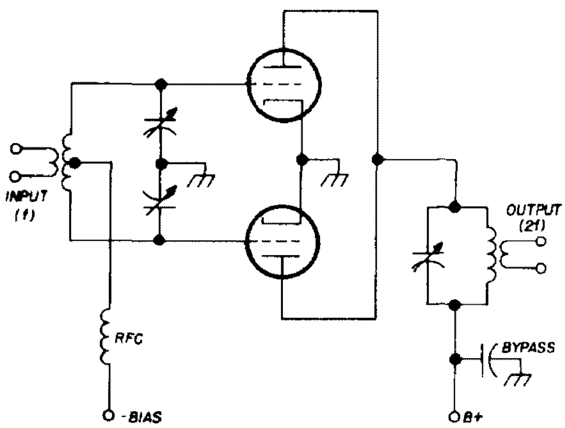


Fig. 2. Vacuum-tube push-push doubler circuit.

transistor multipliers

With the development of the bipolar transistor a new type of multiplier became available – perhaps two new types. The most obvious way to use a transistor as a multiplier is to rely on the nonlinear characteristic of the base-emitter diode. The transistor doubler circuit in **fig. 3** has pi-networks on both input and output.[2]

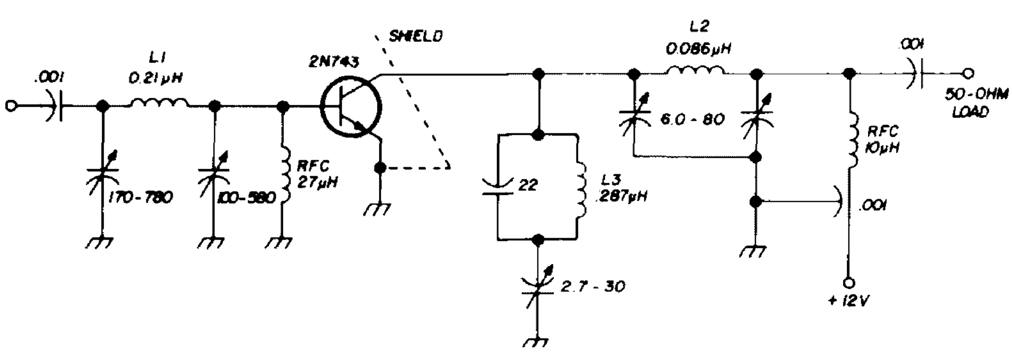


fig. 3. Bipolar-transistor 21- to 42-MHz doubler. L1 is 3 turns no. 516 Air Dux; L2 is 4 turns no. 416 Air Dux; L3 is 7 turns no. 416 Air Dux.

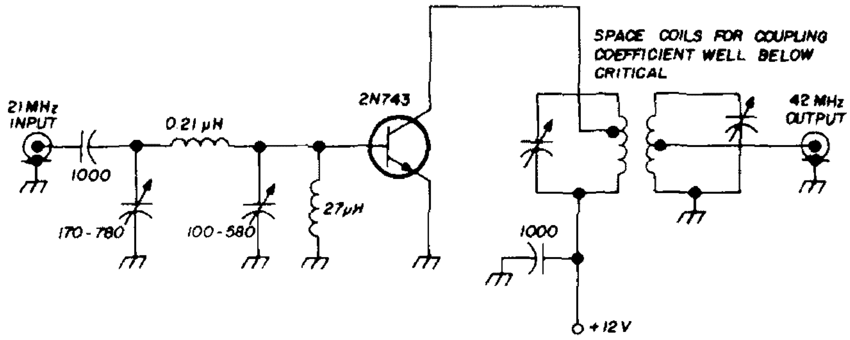


fig. 4. Frequency doubler circuit using a bandpass output circuit.

Since the circuit is a frequency doubler and a pi network is a *lowpass* network, it is necessary to add a fundamental trap consisting of L_3, C_4 and C_5. (If it were a tripler circuit it would be necessary to have both fundamental and second-harmonic traps.) If you use a bandpass coupling network such as a double-tuned circuit these traps may be eliminated. Such a circuit is shown in **fig. 4.** In fact, if you are not too fussy about the rejection of fundamental and unwanted adjacent harmonics the circuit of **fig. 5** can be used. However, **fig. 5** is best suited for doubling because frequency doubling gives the greatest spacing (percentage) of unwanted frequencies.

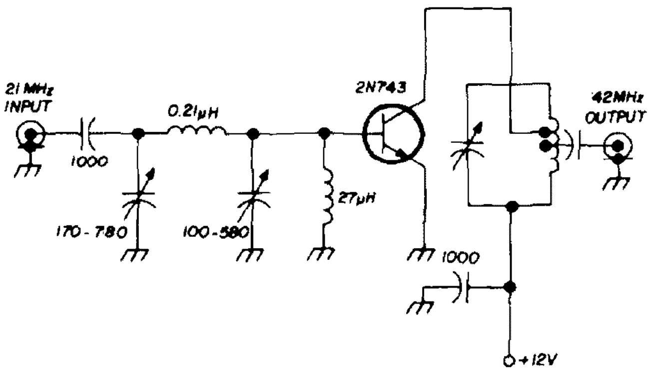
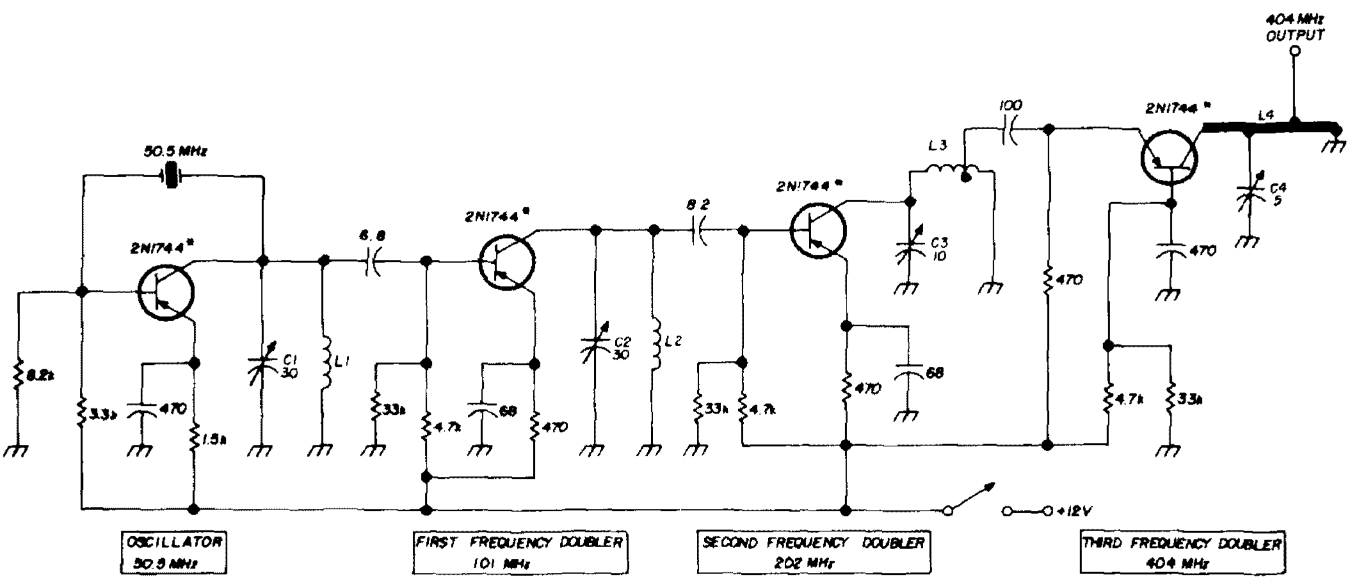


fig. 5. Frequency doubler circuit using a single parallel-tuned output circuit.

Since the transistor does not conduct current until the base-emitter junction is forward biased it is necessary to have at least a few tenths of a volt across this junction for appreciable multiplier action. For this reason, if the drive level is low, it is advisable to use a little dc forward bias to cause the transistor to conduct. However, forward bias and low drive voltage minimize the nonlinearity of the base-



*CASES GROUNDED TO CHASSIS

L1 10 turns no. 18, ½" ID, 5/8" long

L2 3 turns no. 18, ½" ID, ½" long

L3 5 turns no. 18, ¼" ID, ½" long, tap at 1¼ turn from ground

L4 ½" wide copper trap, 2" long, spaced ½" from chassis, tapped ½" from ground

fig. 6. Three frequency doublers which use forward bias of the base-emitter junction because of low-level drive.

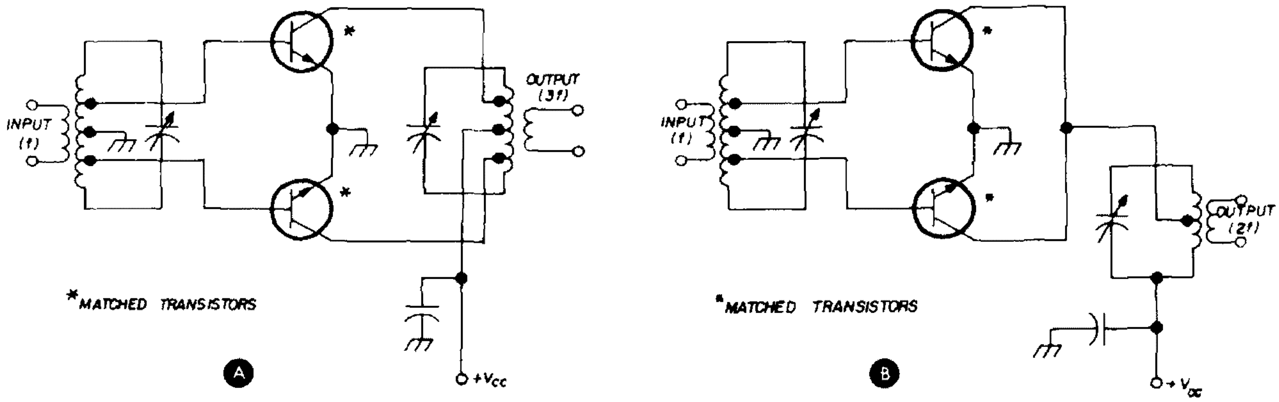


fig. 8. Push-pull tripler circuit is shown in (A). Circuit in (B) is a push-push doubler.

emitter junction and decrease the transistor's effectiveness as a multiplier. A transistorized crystal oscillator and multiplier chain using forward-biased frequency couplers is shown in **fig. 6;** it was originally designed as a local-oscillator source for a 432-MHz converter.[3] Of course, the base-emitter voltage required to forward-bias germanium transistors (as in **fig. 3** through **6**) is lower than that required for silicon transistors.

In a recent article on transistor multipliers W6AJF suggested a number of practical circuit innovations that make frequency multipliers more efficient.[4] One point is to reduce the affect of base-to-collector capacitance. He suggests a couple of ways of doing this: adding a series trap from the base to ground tuned to the output frequency, or using a capacitive divider in the input tuned circuit, making sure that C2 is no larger than -j10 ohms at the output frequency (see **fig. 7B**). W6AJF also suggests the use of a "grid-leak" arrangement to reduce danger of transistor breakdown where the

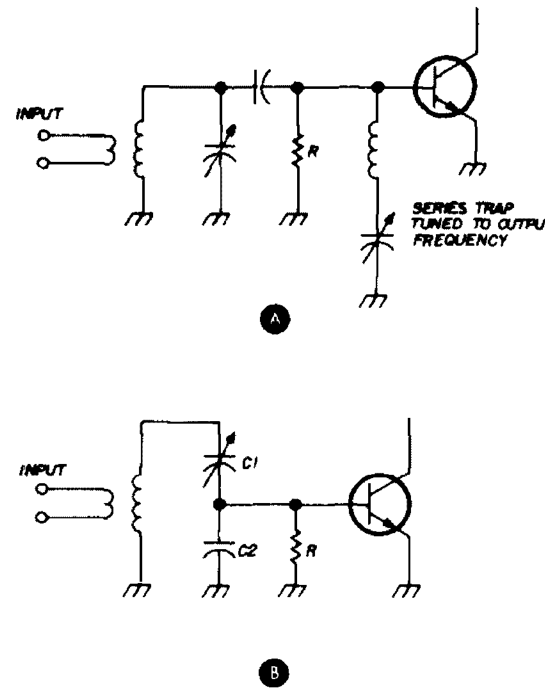


fig. 7. Improving multiplier efficiency: Input of circuit (A) uses a series trap tuned to the output frequency; circuit in (B) uses a large value at C2.

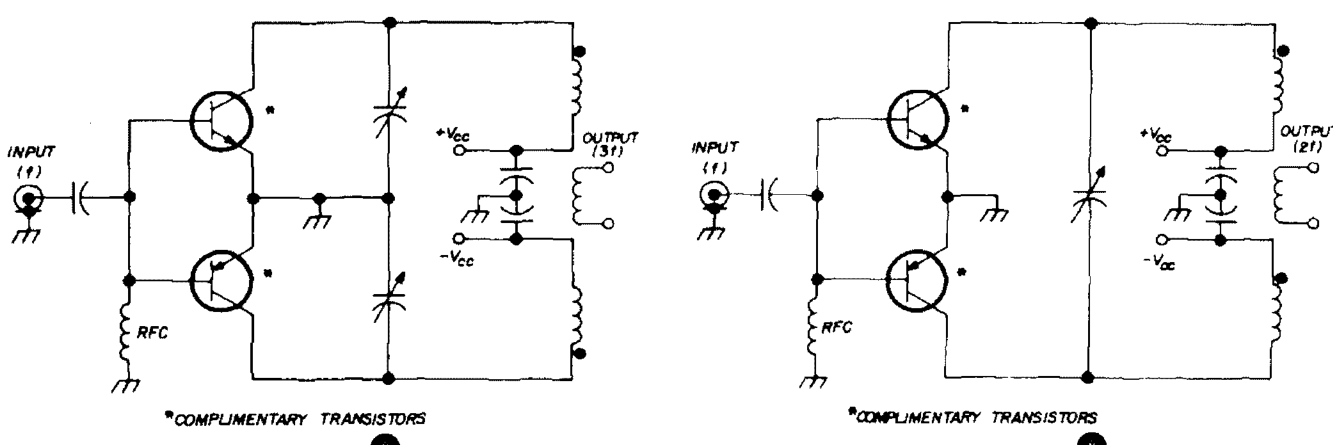


fig. 9. Complimentary transistor frequency multipliers. Push-pull tripler circuit in (A); push-push doubler circuit in (B).

drive voltage exceeds the maximum base-emitter reverse breakdown rating.

As with tubes, it is possible to use push-pull transistor multipliers for odd-harmonic generation and the push-push transistor multipliers for even-harmonic generation (circuits are shown in **fig. 8**). Unlike tubes, however, we can build complimentary circuits with transistors. Since both npn and pnp transistors are available it is possible to build both complimentary push-pull and complimentary push-push multipliers (typical circuits are shown in **fig. 9**).

Of course, you don't have to use the common emitter-configuration in transistor multipliers; common-base circuits for the complimentary push-pull and push-push multipliers are shown in **fig. 10.** It should also be mentioned that transistor matching is desirable in both push-pull and push-push circuits. Anti-matching (finding pnp and npn transistors that have equal but opposite polarity characteristics) is desirable for the complimentary push-pull and push-push circuits. *Dual* npn *and* pnp *silicon transistors* are quite commonly available; also, many companies offer separate npn and pnp transistors which are intended for complimentary use (such as the Motorola 2N3903 and 2N3905). In the case of matched npn transistors it is often less expensive to buy an IC array, such as the

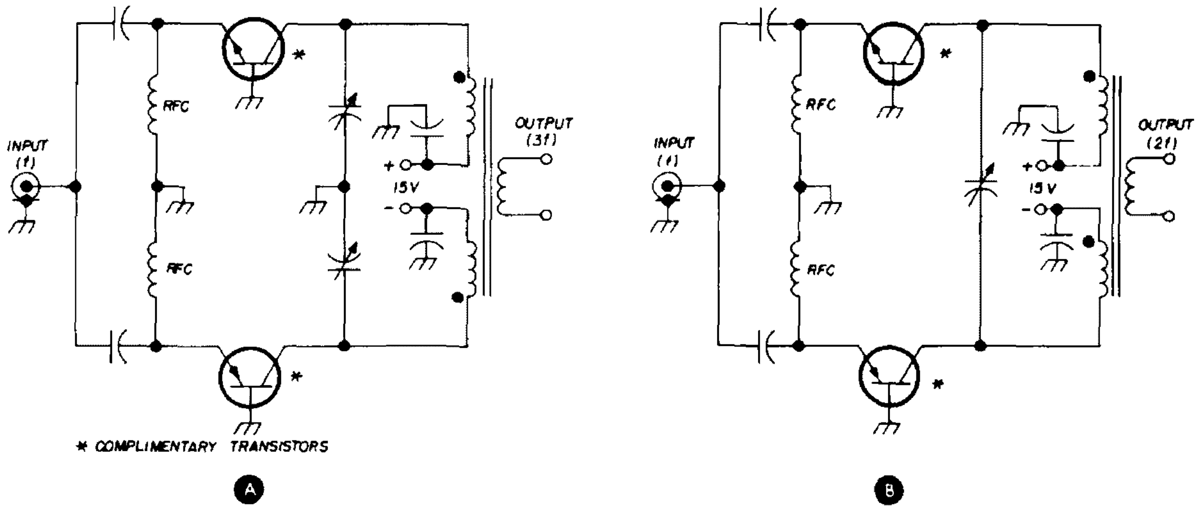


fig. 10. Complimentary common-base multiplier circuits. (A) is push-pull tripler; (B) is push-push doubler.

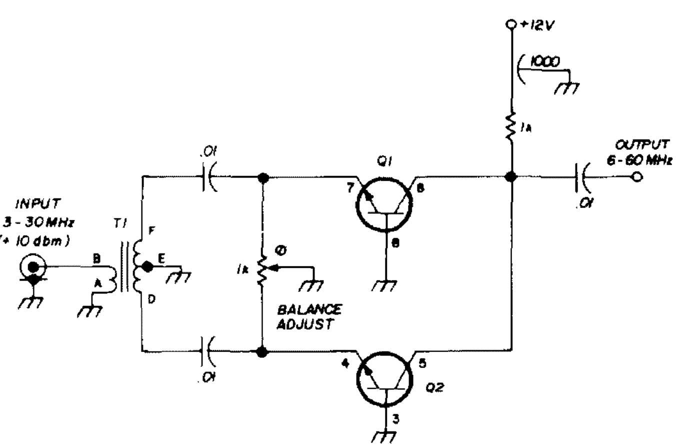


fig. 11. Broadband push-push frequency doubler circuit uses two transistors of a CA3018 IC; numbers indicate IC pin numbers. T1 is a North Hills 50:400-ohm balanced transformer. Ground pin 10 of the CA3018.

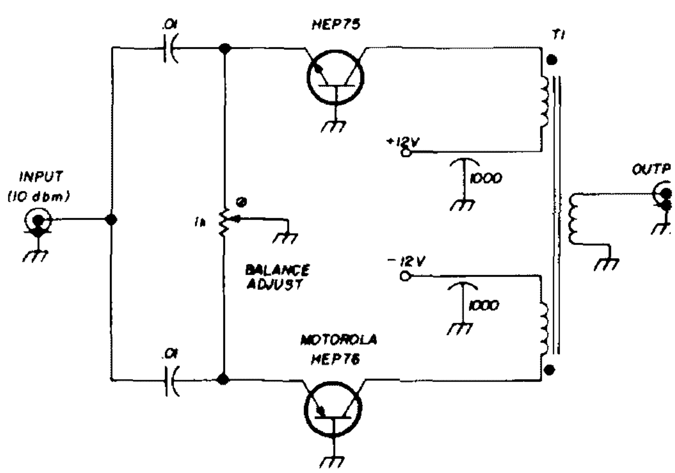


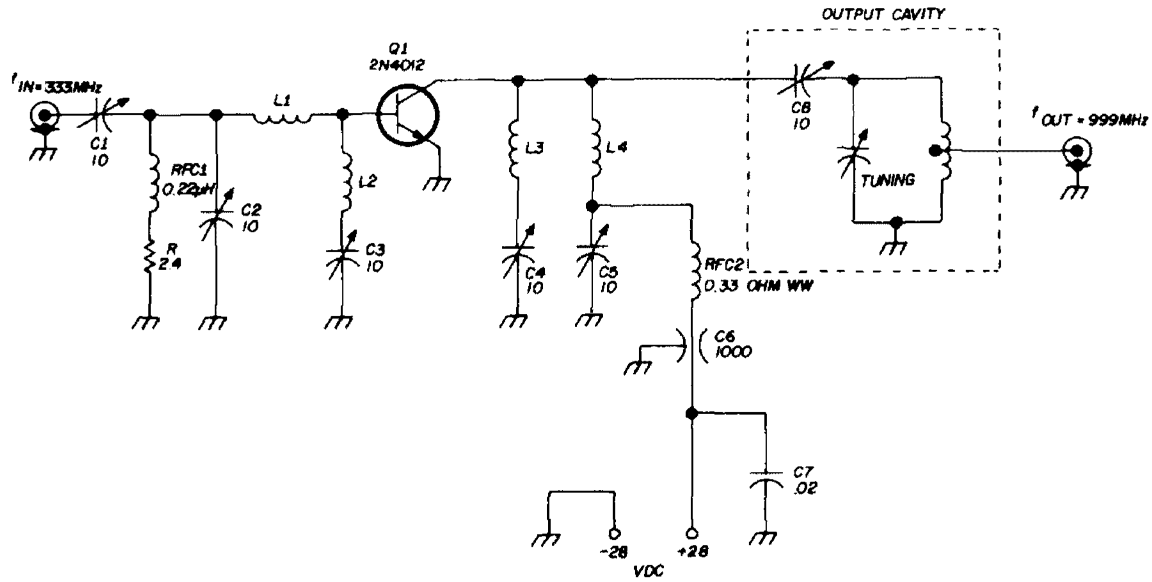
fig. 12. Broadband push-pull complimentary tripler circuit. Primary of T1 is 28 bifilar turns no. 30 on Indiana General CF103-Q1 core; secondary is 2 turns no. 24.

RCA CA3018, than to buy a commercial matched pair; the CA3018 contains 4 matched npn transistors (see **fig. 11**).* A broadband complimentary bipolar tripler is shown in **fig. 12**.

multiplier for tripling from 333 to 999 MHz is shown in **fig. 13**.

fet multipliers

The field-effect transistor is a more



L1	2 turns no. 18, ¼" diameter	L3	2 turns no. 18, 3/8" diameter
L2	1/16-inch wide copper strap, 4" long	L4	1 turn no. 18, ¼" diameter

fig. 13. Uhf transistor parametric multiplier. RFC2 is 0.33-ohm wirewound resistor.

parametric multiplication

A second mechanism that may be used for frequency multiplication with transistors uses the base-collector depletion capacitance.[5, 6] This is called parametric multiplication. As with parametric diode multipliers you must design carefully to obtain anything like optimum efficiency. The parametric effect is a high-level one and only becomes important when the rf voltage is swinging a significant percentage of the collector-to-base voltage. To make the most of the parametric effect the collector circuit must have a number of idler circuits which increase efficiency by reflecting undesired harmonics back to the collector-base capacitance. A typical parametric transistor

recent addition to the amateur's bag of semiconductor tricks. Fets have dropped in price with the availability of plastic packaging and modern production methods, and a number of respectable devices are available for a dollar or less. Fets can be grouped into two general types: Junction and insulated-gate. Both n- and p-channel fets are available in junction and insulated-gate versions.

The fet frequency multiplier can be treated in much the same way as the

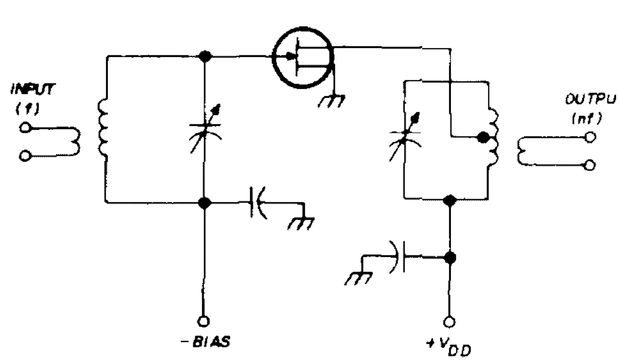


fig. 14. Simple fet frequency miltiplier.

*The broadband transformer used in fig. 11, as well as the one used in fig. 18, is available for $5.50 postpaid from Hank Olson, Post Office Box 339, Menlo Parl, California 94025.

vacuum-tube multiplier, with cutoff bias applied to the gate and rf drive voltage to set the conduction angle indicated in table 1. Two such circuits are shown in **fig. 14.** Both of the fets in **fig. 14** are n-channel devices since most fets for rf are n-channel types, whether junction or insulated-gate types.

Since both n- and p-channel fets are available it is possible to build the complimentary push-pull and push-push multipliers in much the same way as with bipolar transistors. Circuits are shown in **fig. 15.** Conventional push-pull and push-push multipliers are feasible too; examples are shown in **fig. 16.** Matched fet pairs are commonly available, and push-pull and push-push circuits will work best with them. A complete 14-MHz crystal oscillator and fet push-push doubler with 28 MHz output is shown in **fig. 17.** The 28-MHz output is remarkably free of 14 MHz energy due in part to circuit balance, and in part to the resonant circuit, T2. Another fet push-push doubler, one that will work over the entire high-frequency spectrum, is shown in **fig. 18.** A broad-band complimentary fet doubler is in **fig. 19.**

IC multipliers

Two similar devices that have recently been introduced to the IC market are the multiplier and the balanced modulator; either can serve as a frequency doubler. The Motorola MC1595 and MC1596, or their Fairchild equivalents, μA795 and μA796, can be used as doublers by introducing rf drive into *both* inputs. Since the MC1595 and μA795 are *true* multipliers the input signal is multiplied times itself, producing a $\sin^2\theta$ waveform. The $\sin^2\theta$ waveform looks very much like a full-wave rectified sine wave and contains only the second harmonic plus a dc component.*

The MC1596 and μA796 balanced modulators operate in much the same

*If you are accustomed to trigonometric identities, the following equation provides an explanation for multiplier operation:

$$\sin^2\theta = ½(1 - \cos 2\theta)$$

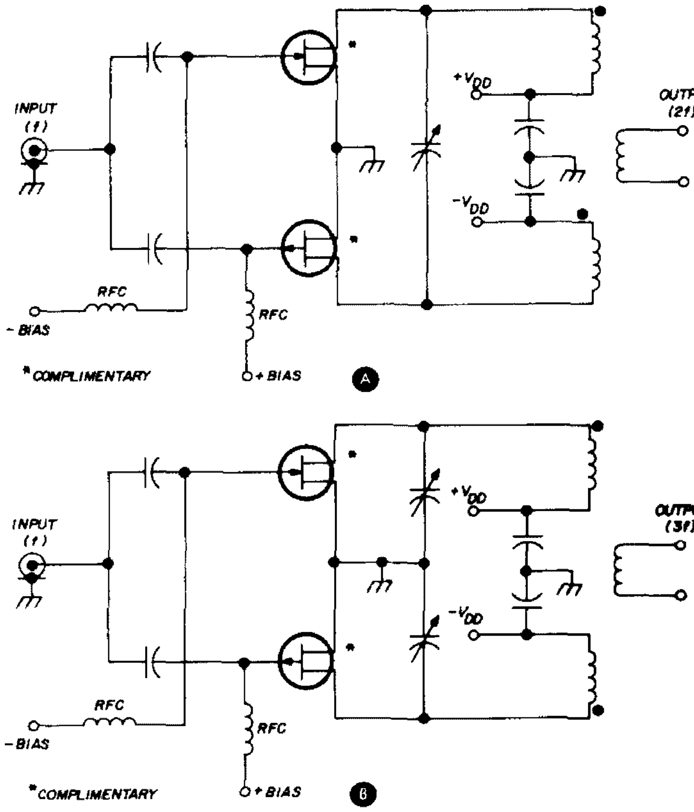


fig. 15. Complimentary fet multiplier circuits. (A) is doubler; (B) is tripler. Fets are matched complimentary pairs.

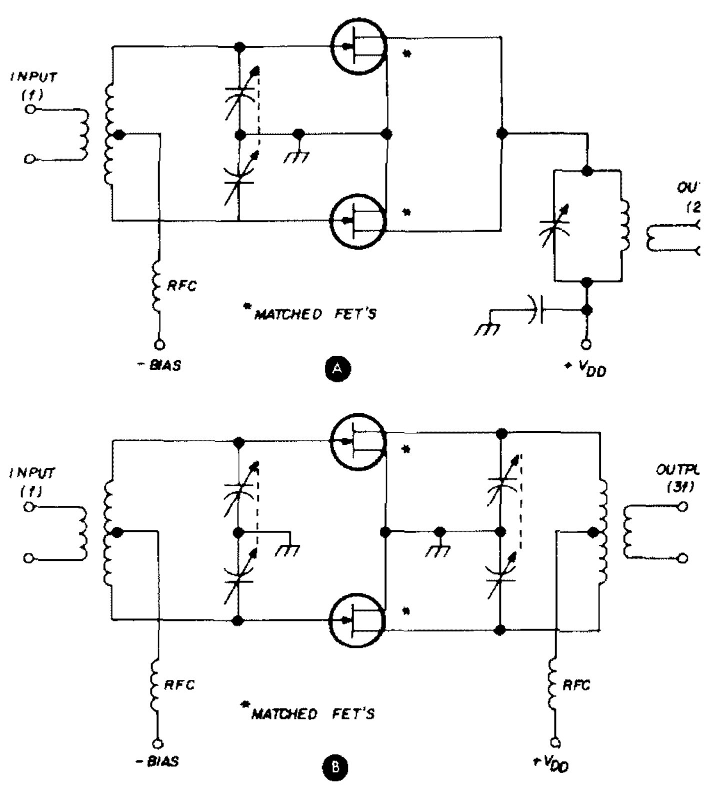


fig. 16. Push-push fet doubler (A), and push-pull fet tripler (B).

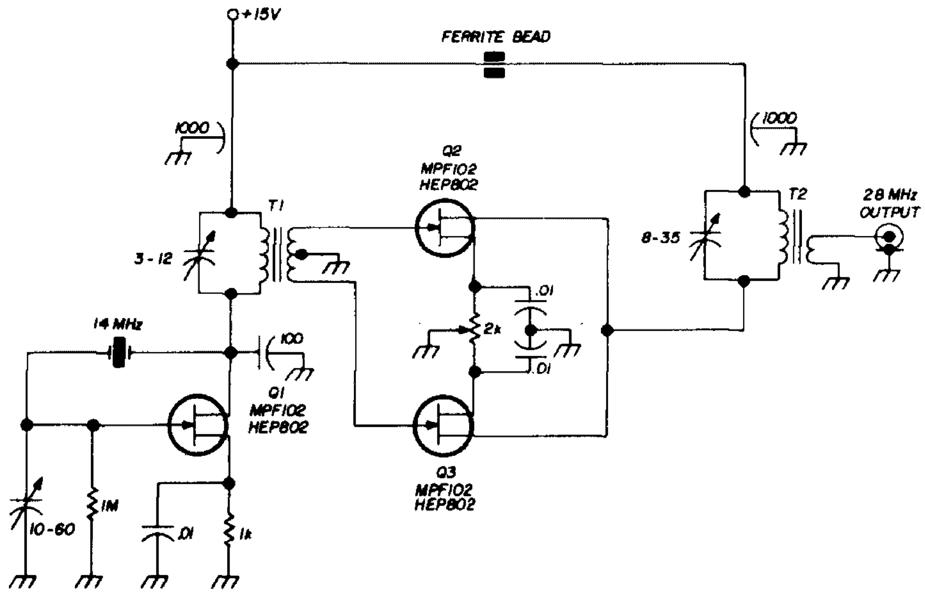


fig. 17. Fet 14-MHz crystal oscillator and push-push doubler. Primary of T1 is 20 turns no. 26 on Amidon T44-6 toroid core; secondary is 4 turns no. 22, center-tapped. Primary of T2 is 13 turns no. 20, on Amidon T44-6 core; secondary is 2 turns no. 20 Fets Q2 and Q3 are selected for approximately equal I_{dss}.

way as the IC multipliers although they are not true multipliers. Frequency-multiplier circuits for the MC1595 and MC1596 are shown in **fig. 20.** Note that the circuit of **fig. 20A** does not depend on tuned circuits to suppress undesired harmonics but does so by circuit balance alone.

diode multipliers

The point-contact diode has been used as a frequency multiplier since the first 1N34s became available, and possibly earlier than that.[7] These diodes were

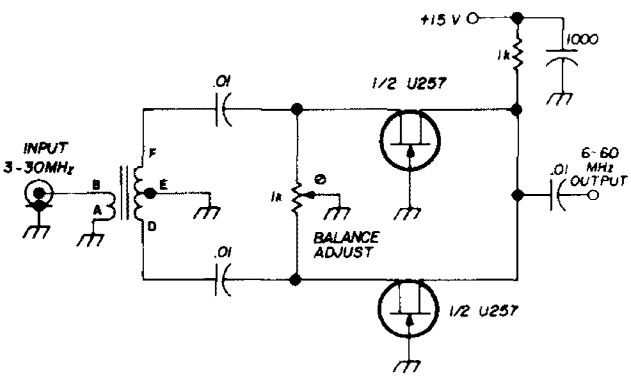


fig. 18. Broadband fet push-push doubler. T1 is broadband North Hills 50:400-ohm balanced transformer. U257 is dual matched Siliconix fet.

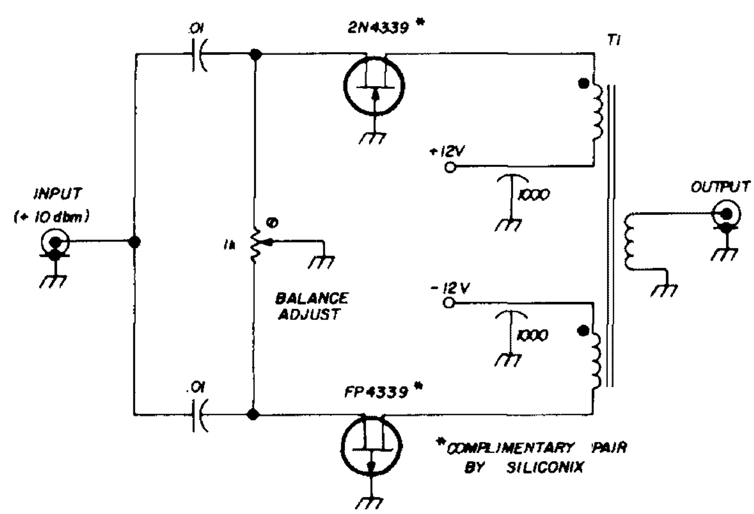


fig. 19. Complimentary fet broadband push-push doubler. Primary of T1 is 28 turns no. 30 bifilar wound on Indiana General CF103-Q1 core; secondary is 2 turns no. 24.

usually used to enhance the harmonics of 100-kHz crystal calibrators, so they cannot be considered to be multipliers of any significant efficiency. Modern versions of such a calibrator would probably still use a 1N100 or 1N270 germanium point-contact diode since these are superior even to good silicon computer diodes.

The hot-carrier or Schottky-barrier diode, such as the Hewlett-Packard HP 5082-2800, is better than either the point-contact germanium or the silicon junction diode. If you use two matched hot-carrier diodes, arranged as a full-wave rectifier, a fairly efficient doubler can be built; a broadband doubler is shown in **fig. 21.** This circuit is very useful as a signal generator accessory.

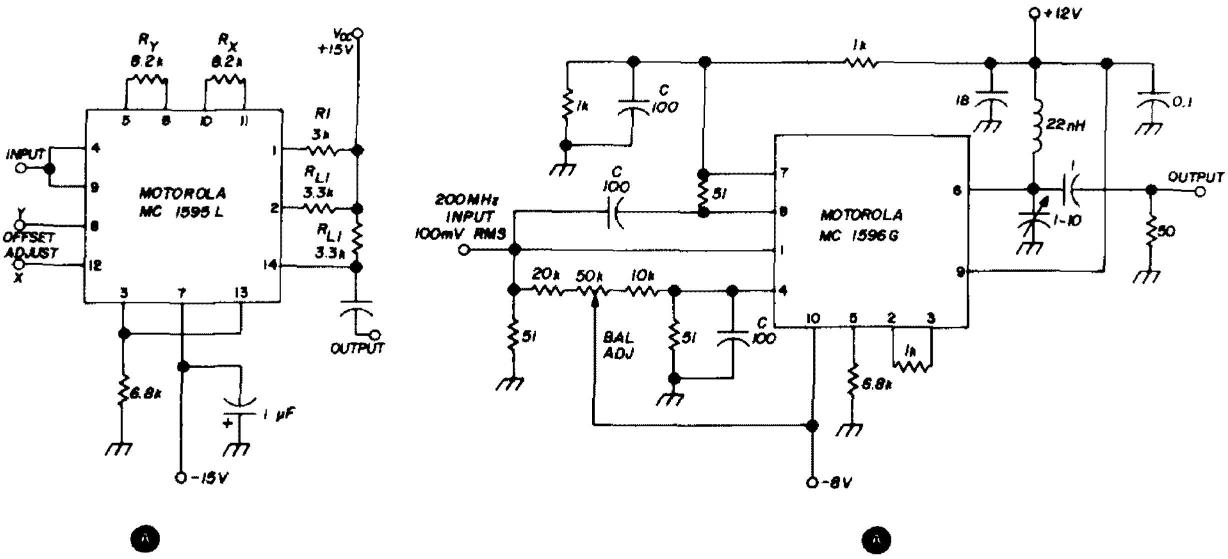


fig. 20 Broadband integrated-circuit frequency doubler is shown in (A). 200-MHz frequency doubler using the Motorola MC1596G in (B). Fairchild uA795/uA796 or Signetics S5995/S5596 may be used in these circuits.

In fact, broadband doublers are sold by Hewlett-Packard and other test equipment manufacturers for doubling the high-frequency output range of their signal generators. The diode doublers put out 12 to 16 dB less second harmonic than input.

Another diode frequency multiplier is the parametric diode multiplier. Many circuits have been published. Like the parametric transistor multiplier, which they pre-date, operation of the parametric diode multiplier relies on the

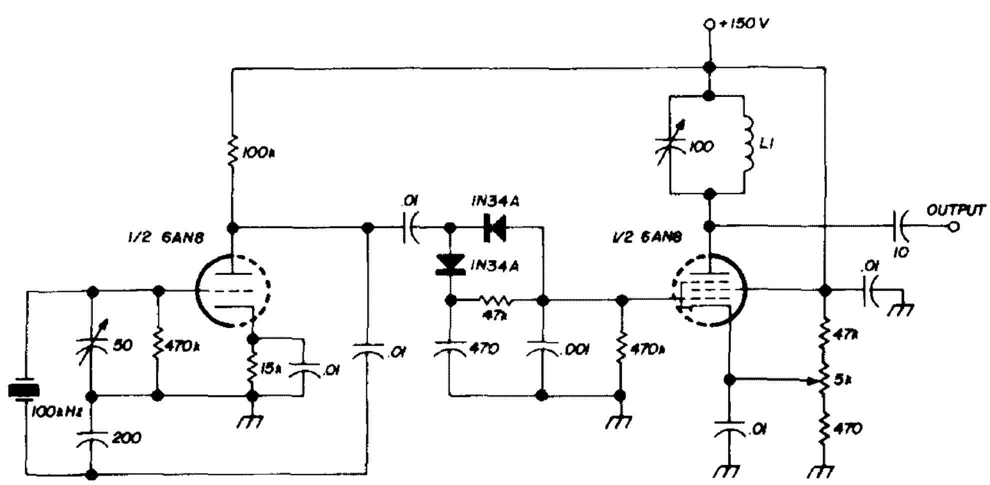
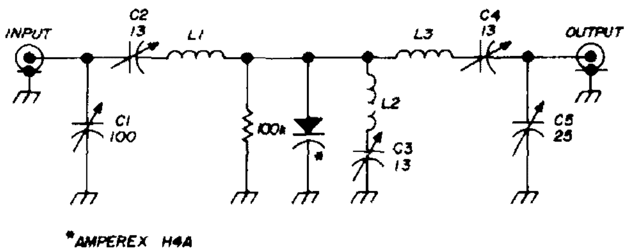


fig. 21. 100-kHz crystal calibrator using germanium point-contact diodes to enhance harmonics. L1 is tuned to the amateur band of interest.

variable capacitance of a reverse-biased junction. Idler circuits are required, and the design and tuneup procedures are extremely tedious. A typical 150 MHz to



L1	6½ turns no. 16, 5/16" diameter, 9/16" long
L2	2 turns no. 12, ¼" diameter, 5/16" long
L3	¼-inch wide copper strap, 1" long, spaced 9/16" from chassis

fig. 22. Medium-power 150- to 450-MHz tripler.

450 MHz diode parametric multiplier is shown in **fig. 22.** This single-ended design is the form most used in amateur equipment.

It is also possible to build a push-pull parametric multiplier which will have the same even-harmonic cancelling properties of the push-pull multipliers I've discussed previously (**fig. 23**). It follows, then, that a push-push parametric multiplier which cancels fundamental and odd harmonics could also be built. Such a circuit is shown in **fig. 24.**

A later development in the way of multiplier diodes is the step-recovery diode. Unlike the parametric diode or varactor, the step-recovery diode does not depend principally on the nonlinearity of its depletion capacitance for harmonic generation. Instead, it depends on the fact that when it is *forward* biased a charge is stored; when the diode is subsequently reverse biased the charge is dumped. In a properly designed circuit the step-recovery diode will be driven into forward conduction by positive excursions of the fundamental drive voltage; the stored charge will dump into a tuned circuit at the desired output frequency.

In a step-recovery frequency multiplier the harmonic power is proportional to $1/n$; where n is the harmonic number. This is much better than the power falloff of a varactor multiplier where harmonic power is proportional to $1/n^2$. In a 144 MHz to 432 MHz tripler, for example, harmonic power would be 3 times greater with a step-recovery multiplier circuit. **Fig. 25** shows efficiency vs harmonic number n for a typical step-recovery diode multiplier. Note that large values of n are quite feasible.[8]

There's another advantage, too: Only the desired output frequency is tuned in a step-recovery diode multiplier – no elaborate idler system is required.

Until recently step-recovery diodes

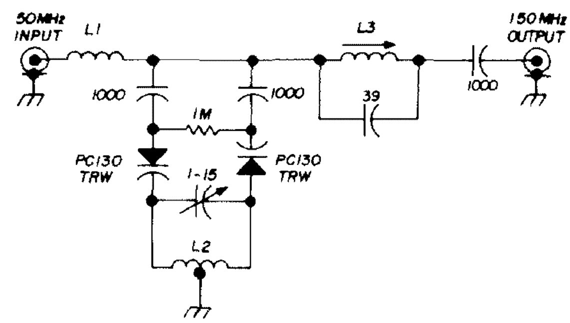


fig. 23. Balanced parametric diode 50- to 150-MHz tripler features conversion efficiency of 70%. L1 is 0.18 µH choke. L2 is 7 turns no. 18, 5/16" diameter, 3/4" long, center-tapped. L3 is 8 turns no. 20 closewound on ½-inch form.

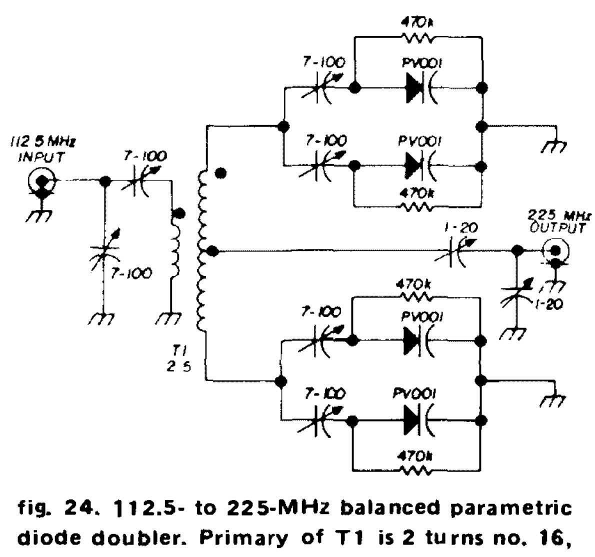


fig. 24. 112.5- to 225-MHz balanced parametric diode doubler. Primary of T1 is 2 turns no. 16, 3/8" diameter, 1-1/2" long, center-tapped; secondary is 2½ turns no. 16 on each side of primary. Conversion efficiency is 75%. Varactors are TRW types.

have been very high-priced and available only to industry. However, one low-cost step-recovery diode is available for amateur experiments. The Siliconix SV110 is available in a standard axial-lead diode package for $5.50.*

Fig. 26 shows the general circuit arrangement for a step-recovery diode frequency multiplier. With the Siliconix SV110 step-recovery diode conversion efficiency falls off as $\frac{200\%}{n}$ up to 3000 MHz so you can expect maximum efficiency of about 22% in a 144 MHz to 1296 MHz times-9 multiplier. To realize anywhere near this theoretical efficiency, however, circuit losses must be low.

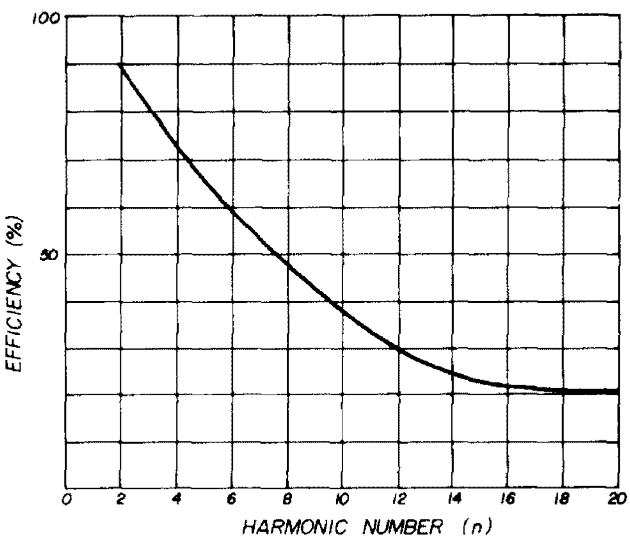


fig. 25. Typical output efficiency of a step-recovery diode multiplier as a function of the harmonic number, n.

Note the similarity between step-recovery diode multipliers and parametric diode multipliers. The circuits are quite similar, so it is also possible to build balanced step-recovery diode multipliers which cancel either odd or even harmonics. However, the advantages of balanced step-recovery diode multipliers is decreased because no idler circuits are eliminated.

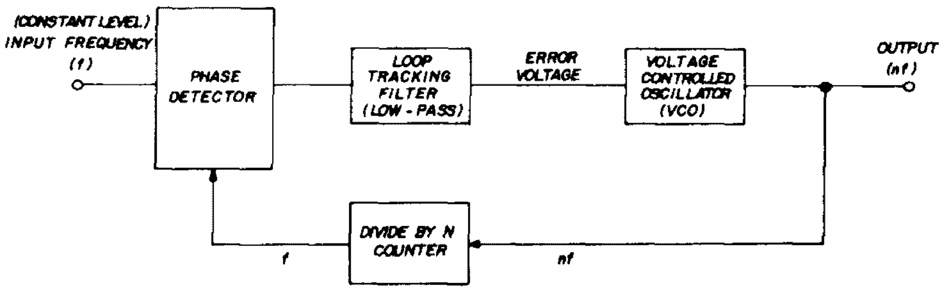


fig. 27. Basic elements of a phase-locked loop frequency multiplier.

phase-locked multipliers

In recent years another, a more complex method of frequency multiplication has been used for special purposes; this is the phase-locked loop frequency multiplier. Frequency synthesizers often make use of this technique because of its inherent signal cleanliness. The system effectively multiplies, but circuit operation is actually accomplished by frequency division. The general nature of the system is shown in fig. 27. To use this system you build the oscillator to the same frequency as the desired output, then phase lock it to a subharmonic. If

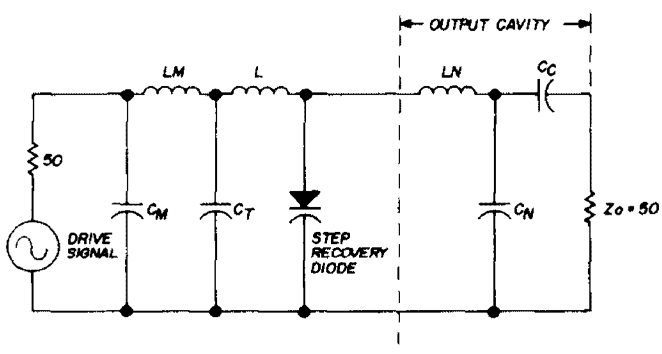


fig. 26. Generalized circuit of a step-recovery diode frequency multiplier.

*$5.50 from Siliconix Incorporated, 1140 West Evelyn Avenue, Sunnyvale, California 94086.

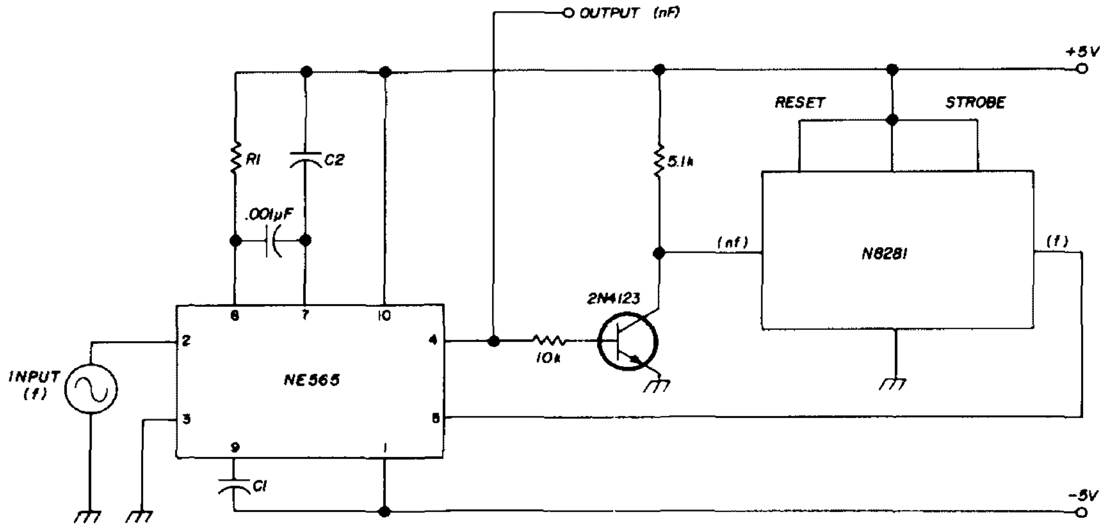


fig. 28. Phase-locked loop multiplier for output frequencies up to 500 kHz uses Signetics NE565.

the oscillator is voltage-controlled the only input to it is the dc *error voltage,* so the output can be made nearly perfectly clean (since no rf is fed into the vco).

Since the phase-locked loop technique is rather complex it has been used only in expensive electronic systems. However, now the phase-locked loop has been tucked into an IC (or two). Because of the advance of IC technology it is a fairly simple matter to build a phase-locked loop frequency multiplier. **Fig. 28** shows a Signetics NE565 phase-locked loop with a divide-by-n counter; in this case a four-stage binary counter, the Signetics N8281. It is not necessary to use this particular counter as nearly any TTL counter or combination of flip-flops will do. Since flip-flops may be connected in a variety of ways to divide by nearly any number almost any harmonic can be multiplied by this system. The Signetics NE565 has an upper frequency limit of 500 kHz, but the Signetics NE562 may be used to 60 MHz as shown in **fig. 29.**

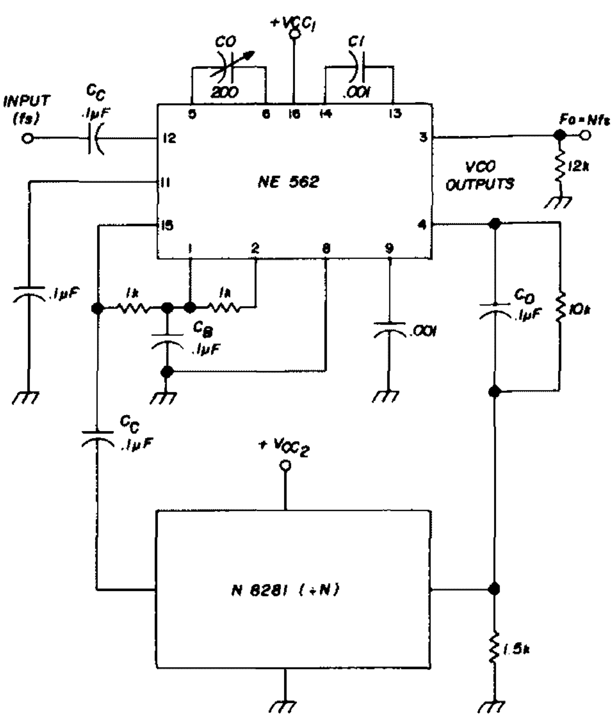


fig. 29. Signetics NE562 allows frequency multiplier operation up to 60 MHz (output). Capacitor C1 is a low-pass filter capacitor; Co sets frequency of operation.

references

1. F. E. Terman, "Radio Engineering," 3rd edition, McGraw-Hill, New York, 1947, page 394.
2. Texas Instruments, "Transistor Circuit Design," McGraw-Hill, New York, 1963, page 328.
3. Philco, "Application Lab Report No. 711," June, 1961.
4. F. C. Jones, W6AJF, "Transistor Frequency Multipliers," *ham radio,* June, 1970, page 49.
5. H. D. Olson, W6GXN, "VHF Parametric Transistor Multipliers," *73,* July, 1966, page 20.
6. J. Cochrane, "Transistor Multiplier versus Transistor-Varactor Multiplier," Motorola application note AN-243, August, 1966.
7. "The Radio Amateur's Handbook," 35th edition, A. R. R. L., 1958, page 517.
8. "Harmonic Generation using Step-Recovery Diodes and SRD Modules," Hewlett-Packard application note 920.
9. "Frequency Multiplication with the Step-Recovery Diode," Hewlett-Packard application note 908.

ham radio

rf clipper for the Collins S-line

This rf speech-processing circuit offers increased talk power without distortion or splatter — circuit can be adapted to other transmitters

Bruce Clark, K6JYO, 1019 El Dorado Drive, Fullerton, California 92632

Rf speech clipping is the most effective technique for increasing the average to peak power output of a ssb transmitter without increasing the transmitted bandwidth or in-band distortion products. The high effectiveness of rf clipping was dramatically illustrated by W6JES.[1] This article describes a highly effective rf speech clipper that is installed in a Collins 32S1 transmitter; the same circuit may also be used in the 32S3. The Collins S-line transmitter is one of the best rigs for added rf clipping circuitry because of

1. A beefy power supply, capable of supplying a high average current level.

2. The use of 6146 transmitting-type final-amplifier tubes coupled with an effective rf compression loop (alc) and negative feedback to the final amplifier stage for reduction of intermodulation products.

3. A stable, effective balanced modulator capable of providing carrier suppression in excess of 50 dB for long periods of time.

Because of the highly effective alc loop very little additional performance improvement is gained through the use of audio processing techniques such as compressors and audio clipper/filters. The *apparent* improvement noticed with these devices usually comes with a sacrifice in overall intelligibility because the noise power of the processor-generated distortion products increases nearly as much as the voice power. The net result is a sloppy signal of poor overall intelligibility — exactly the thing we are trying to improve by using a signal processing device.

Quantitatively, an overall improvement of about 1-dB is all you can expect from an audio processor without adding appreciable distortion. Any S-meter reports to the contrary are due to noise and distortion components of the processor that hang the slow decay agc of the receiver and cause a higher average meter reading.

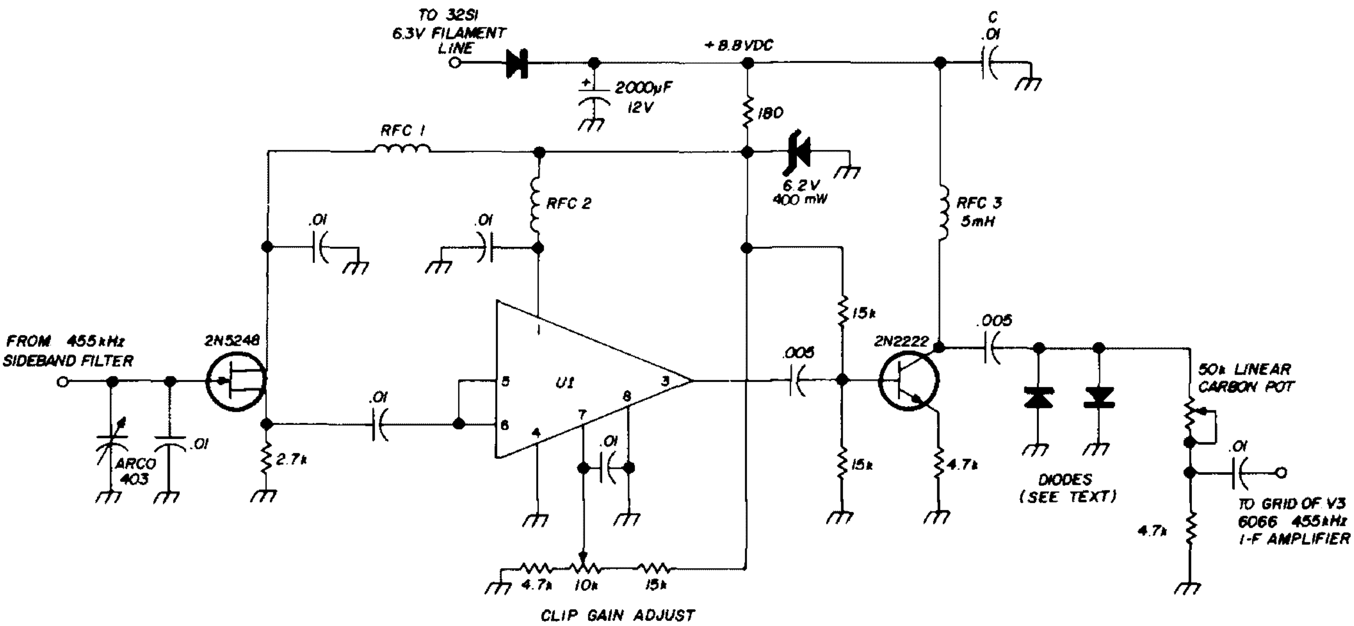
rf clipping

The answer to the signal processor question, at least for the Collins S-line* is rf speech clipping. With rf speech clipping all processing takes place at the 455-kHz i-f. All harmonic distortion components are multiplied in frequency and are no longer close to their audio fundamentals; therefore, they can be removed easily by a bandpass filter. Intermodulation distortion products are still present, but they don't become objectionable at clipping levels up to 20 dB with the S-line transmitter.

clipper circuit

The rf-clipper circuit shown in **fig. 1** is currently installed in my 32S1 transmitter. The circuit consists of an fet source follower, a variable gain IC amplifier, a transistor driver and diode clipper. The 2N5248 input stage matches the high output impedance of the 455-kHz mechanical filter to the IC amplifier. The variable-gain IC amplifier is based on the Plessey SL612C.* (A Motorola MC1590 IC should also work well in this circuit but one was not available when I built the clipper.)

The single transistor power-amplifier stage following the IC amplifier is necessary for maximum clipping level while providing enough surplus gain to allow the clipper to be taken in and out of operation by merely turning the gain control. The clipping diodes are connected across the collector circuit of this amplifier.



C1	pF (Arco 403)	RFC1	7 turns no. 28 enameled on 0.060" diameter
CR1	6.2-volt, ½-watt zener diode	RFC2	Ferroxcube 3K2A ferrite core
CR2,CR3	Hot-carrier diodes, HP2800 (see text)	RFC3	5 mH (J. W. Miller 6304 or 70F-473A1)
CR4	silicon rectifier	U1	Plessey Microelectronics SL612C

fig. 1. Schematic for an rf signal clipper designed for the Collins S-line. Circuit can be adapted to other filter-type ssb transmitters which have sufficient power-supply reserve.

Several types of diodes were tried in this circuit including 1N914 silicon, 1N277 germanium and HP2800 hot-carrier diodes. As was expected at this frequency all types worked well. However, silicon devices have a higher conduction threshold in the forward direction, so require more clipper gain for a particular clipping level. The leakage of germanium diodes increases markedly with temperature, causing a noticeably lower level of clipping with a softness or rounding of the clipped peaks.

This leaves the hot-carrier diode as the best overall choice. The price of hot-carrier diodes is now less than $1.00 from your local Hewlett-Packard regional of-

*The SL612C integrated circuit is $5.65 from Plessey Electronics Corporation, 170 Finn Court, Farmingdale, Long Island, New York 11735.

fice or from Hal Devices.* Matched pairs and quads are also available from both sources. However at 455 kHz, the characteristics of individual devices are close enough for very symmetrical clipping of the ssb envelope. A resistive attenuator across the output of the clipper reduces the output level to a suitable value for the 6DC6 i-f amplifier stage. The adjustable attenuator compensates for gain variations in the devices.

The alternative approach is to use the original filter in the sideband filter location, and use the second filter for crud rejection. The second filter may have considerably wider bandwidth than the sideband filter without causing any serious increase in distortion products.[2]

construction

My rf clipper is built on a 2x3-inch piece of copper-clad perforated Vector

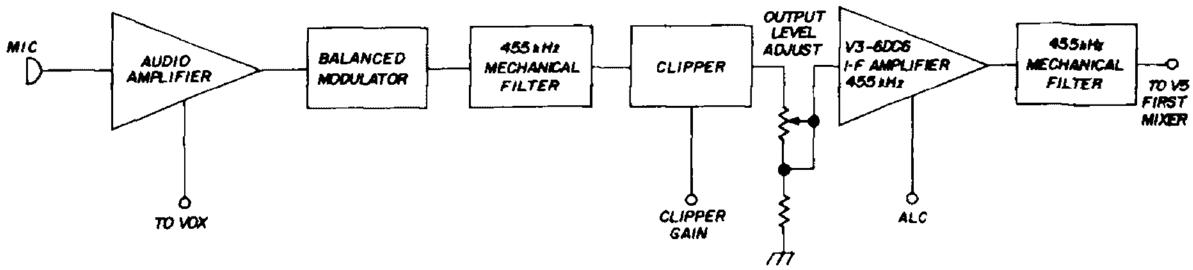


fig. 2. Rf clipper placement in the S-line transmitter. Circuit requires additional 455-kHz mechanical filter.

installation

Fig. 2 shows the placement of the rf clipper circuit in the 32S1 transmitter. This arrangement was arrived at after much experimentation. It results in a stable unit with a minimum of wiring modifications to the 32S1 circuitry and no holes. Two mechanical filters are used. The first filter suppresses the unwanted side-band. The second filter blocks the unwanted harmonic distortion and out-of-band intermodulation-distortion products that are generated by the clipper.

I found it easiest to leave the original 455-kHz filter between the 6DC6 i-f amplifier (V3) and the 12AT7 first mixer (V5). This filter now serves as the crud rejection filter. I obtained a second 2.1-kHz filter for use as the sideband-suppression filter; I used a Collins F455FA21. The carrier crystal requirements were very close to those of the original ssb filter. This may be difficult to achieve if you buy a filter from a different manufacturer.

*HAL Devices, Box 365H, Urbana, Illinois 61801.

board. A Vector pad-cutting tool was used to insulate the holes needed for component mounting. Vector T-28 pins were used for component anchors and for circuit terminations to the power supply, input/output and gain-control pot.

The board is installed underneath the 32S1 chassis next to the existing mechanical filter. The wires for the remote gain control are routed through the PTO power cable hole to the pot which is mounted on an L-bracket at the rear of the PTO; the bracket is mounted with one of the PTO cover screws. The board is mounted with heavy no. 14 ground leads soldered to the copper-clad board; these leads are attached to solder lugs installed under existing screws in the immediate underchassis area.

adjustment

Adjustment of the rf clipper is simple if you have a monitorscope. If a monitorscope isn't available, beg, borrow or steal one, at least for the initial setup. Using a single tone (32S1 in *tune,* then *lock-key*) tune and load the transmitter in the normal manner. Change the *mode* switch

to either sideband and adjust the *mic* audio gain to the normal level. This should be between 8:30 and 10 o'clock, depending on the type of michrophone you're using.

Change the *meter* switch to *alc* and check alc action. With the rf clipper turned off peaks should be between 4 and 8 dB. If required, adjust the output attenuator pot for this amount of average alc action. Now, advance the clipper gain control while talking; the slope of the scope pattern should change markedly, flattening out as shown in fig. 3. The average wattmeter reading will also increase. No flat-topping of the ssb speech envelope should be detected. Nor should the vertical scope deflection increase beyond the value first noted under single-tone conditions when tuning up with the *function* switch in *lock-key.* The alc level now should increase to 10 to 12 dB on peaks; if it goes above this level reduce audio gain or increase the value of the output attenuator until the correct alc levels are obtained.

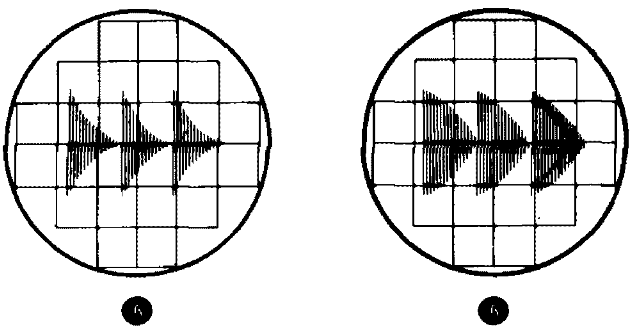


fig. 3. Typical rf envelope patterns. (A) shows unprocessed ssb speech; (B) shows processed speech envelope with about 15-dB clipping.

performance

Performance of the rf clipper unit has been up to expectations. The results agree closely with those predicted in the *QST* article;[1] that is, an improvement of 8 to 9 dB in signal intelligibility along with a 10- to 11-dB gain in average-to-PEP output-power ratio.

When using rf clipping, several precautions must be observed, especially with regard to the additional strain placed on final amplifier and driver due to the large increase in average input power. The resultant higher temperatures can have a detrimental effect on tube and component life, particularly in the area of the final amplifier compartment.

I eliminated part of the problem by substituting 7212/6146W tubes in the final. I also added a Rotron Whisperfan on top of the transmitter cabinet that blows cooling air downward through the box; the fan is simply mounted on rubber grommets and sits on the cabinet top. Power for the fan is obtained from the 516F2 power supply by tapping into the switched ac line. A small 2-pin Jones plug connects the fan power cord to a short 2-inch line coming out of the 516F2 power supply.

You may consider installing 6146B/8298A tubes in the final amplifier. However, the modification necessary to increase the screen voltage is not recommended with the clipper as the power supply and tube dissipation ratings will be grossly exceeded.

I feel that rf clipping is a must for optimum ssb talk power in any transmitter with sufficient reserve power supply capability. For example, this circuit can be adapted easily for use in transmitters using crystal filters such as the Heath SB400; only minor changes in the crystal filter impedance-matching circuits should be required. Use of the unit for about six months has resulted in a marked reduction of TCIP – "time calling in pileups."

references

1. Collins, W6JES, "Ordinary and Processed Speech in SSB Applications," *QST,* January, 1969, page 17.
2. S. Hudson, ZL1AFO, "Filters for Speech Clipping," *ham radio,* July, 1970, page 74.
3. W. Squires, WZPUL, E. Clegg, W2LOY, "Speech Clipping for Single Sideband," *QST,* July, 1964, page 11.
4. Collins Radio Company, "Principles of Single Sideband," September, 1960.
5. E. W. Pappenfus, et al, "SSB Principles and Circuits," McGraw-Hill, New York, 1964.

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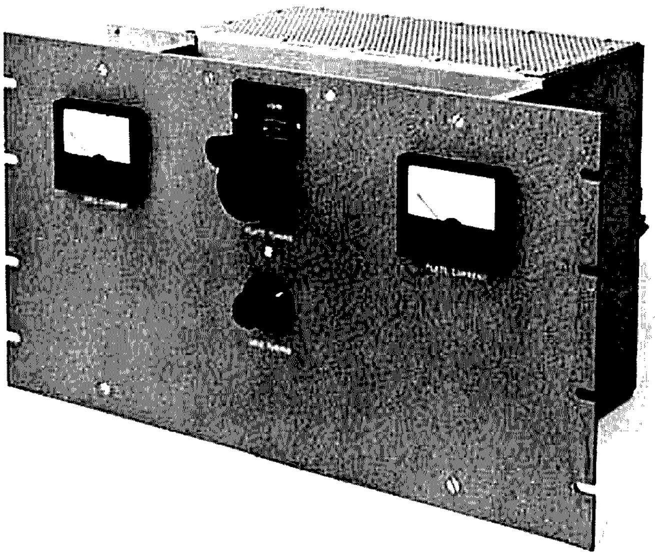


high performance 144-MHz power amplifier

This efficient easy-to-tune grounded-grid 8877 amplifier can be run at 2000 watts PEP ssb or 1000 watts cw

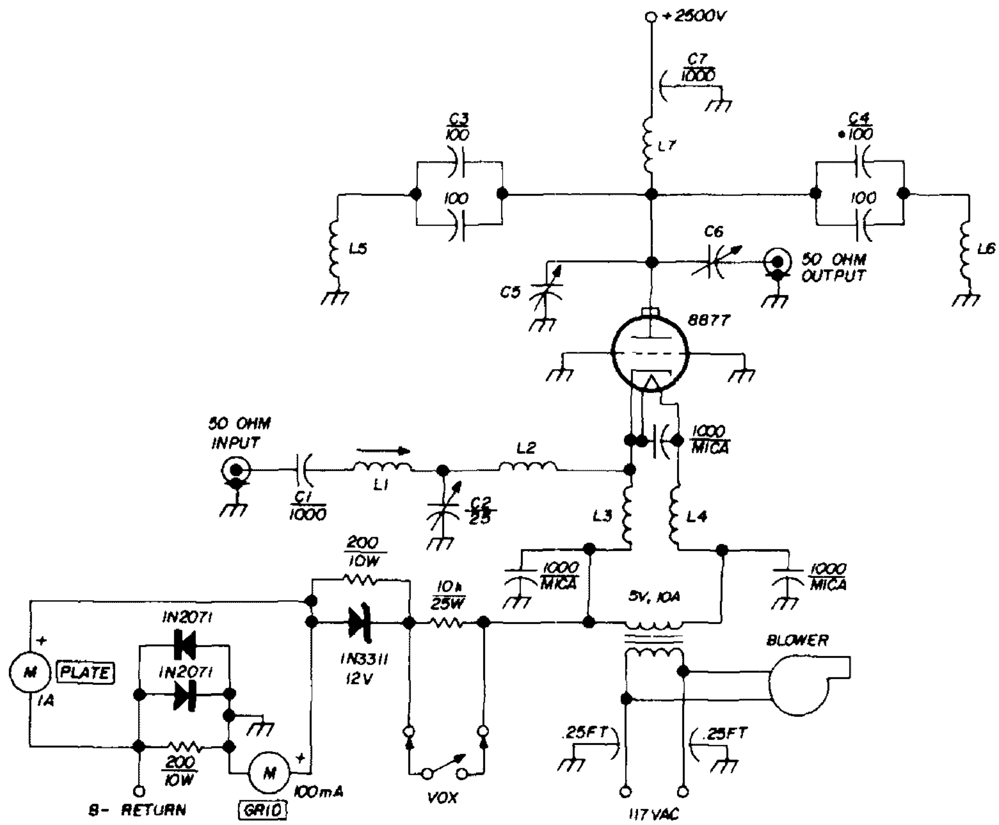
Robert I. Sutherland, W6UOV, EIMAC Division of Varian

The new Eimac 8877 is a ceramic/metal high-mu triode rated for use up to 250 MHz. Operation of this tube at 50 MHz proved to be so satisfactory[1] that other 8877 amplifiers have been designed and built for frequencies up to 350 MHz. Two of these amplifiers are of interest to the serious vhf operator. One amplifier is designed for the amateur 2-meter band and is described here. The other amplifier covers the range from 150 to 230 MHz, and is well suited for use on the amateur 220-MHz band; it will be described later.

The 8877 triode has good division between plate and grid current and low intermodulation distortion. It has a plate dissipation rating of 1500 watts and mu of approximately 200. The cathode is indirectly heated; filament requirements are 5.0 volts at 10 amperes. The tube base mates with a standard septar socket.

This 144-MHz 8877 linear amplifier is designed for the serious DXer who demands reliable service combined with

good linearity and efficiency. The compact grounded-grid design presented here uses a half-wave plate line[2] and a lumped T-network input circuit. The amplifier reserve. For operation at 2000 watts PEP the plate voltage should be between 2500 and 3000 volts; under these conditions the amplifier will deliver 1240 watts



C1	1000 pF (Centralab 858S-1000)
C2	25 pF variable (Hammarlund HFA-25B)
C3,C4	each consists of two parallel-connected 100-pF, 5000-V capacitors (Centralab 850S-100)
C5	plate-tuning capacitor (see fig. 3)
C6	output loading capacitor (see fig. 2)
C7	1000 pF, 4 kV feedthrough (Erie 2498)
L1	5 turns no. 14, 3/4" long on 1/2" diameter form with white tuning slug (CTC 1538-4-3)
L2	4 turns no. 14, air wound, 3/4" long
L3,L4	10 turns no. 12 enameled, bifilar wound, 5/8" diameter
L5,L6	plate resonators (see fig. 4)
L7	7 turns no. 14 wire, 5/8" diameter, 1-3/8" long

fig. 1. Schematic for the grounded-grid two-meter triode amplifier. Operating bias for the 8877 is supplied by a 12-volt zener diode in the cathode lead.

requires no neutralization, is completely stable and free of parasitics, and is very easy to operate.

This amplifier is designed for continuous duty operation at the 1000-watt dc input level, and can develop 2000 watts PEP input for ssb operation with ample output. With the higher plate-voltage supply, up to 13.8-dB gain can be obtained with an amplifier efficiency of 62%.

the circuit

In the amplifier circuit in fig. 1 the 8877 grid is operated at dc ground. The

grid ring at the base of the tube provides a low-inductance path between the grid element and the chassis. Plate and grid currents are measured in the cathode-return lead; a 12-volt, 50-watt zener diode in series with the negative return sets the desired value of idling current. Two additional diodes are shunted across the meter circuit to protect the instruments.

Standby plate current of the 8877 is reduced to a very low value by the 10,000-ohm cathode resistor; this resistor is shorted out when the vox circuit is energized, permitting the tube to operate in normal fashion.

A 200-ohm safety resistor insures that the negative power circuit of the amplifier does not rise above ground potential if the positive side of the plate-voltage supply is accidentally grounded. A second safety resistor across the 1N3311 zener diode prevents the cathode potential from rising if the zener should accidentally burn open.

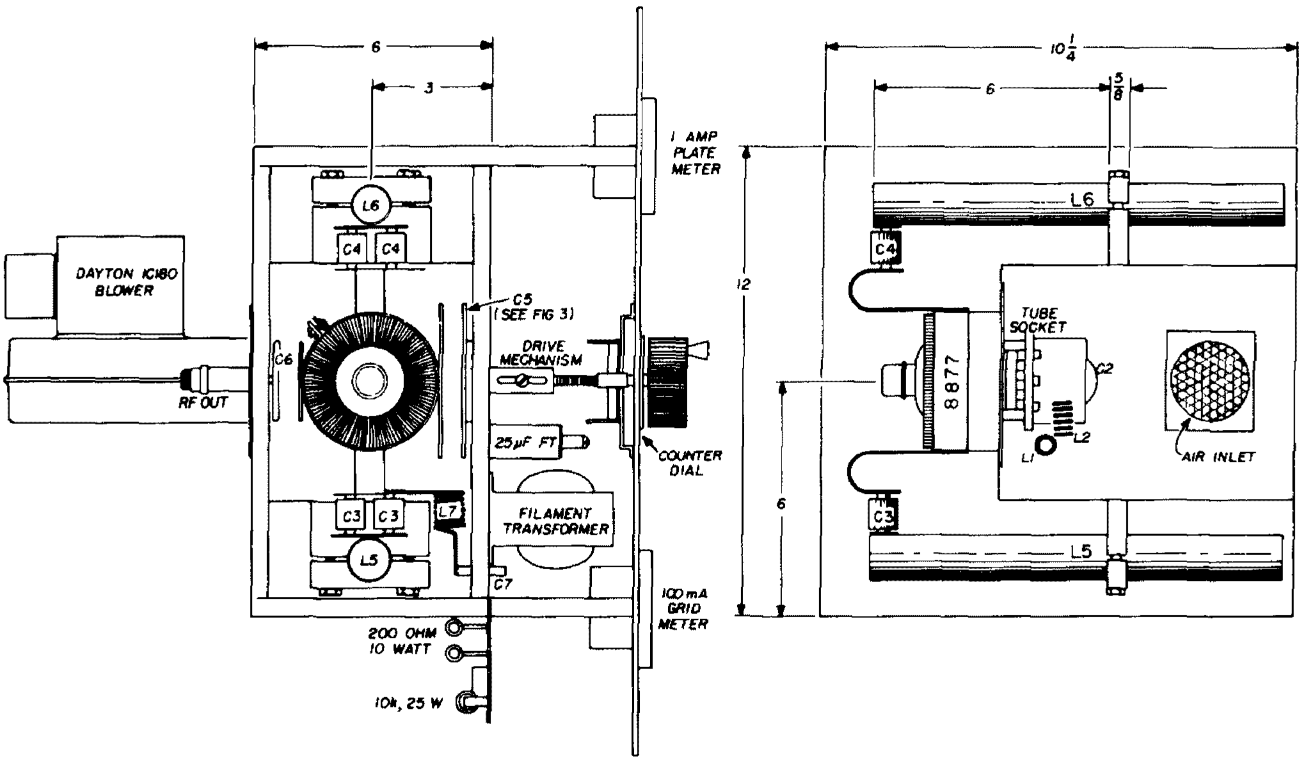


fig. 2. Structural details of the amplifier show the relative size and position of the various components. Assembly is made of aluminum panels.

input circuit

The cathode input matching circuit is a T-network which matches a 50 ohm termination to the input impedance of the tube (about 54 ohms in parallel with 26 pF). The network consists of two series-connected inductors and a shunt capacitor. One inductor and the capacitor are variable so the network is able to cover a wide range of impedance transformations.

The variable inductor (L1) is mounted on the rear wall of the chassis and may be adjusted from the rear of the amplifier. The input tuning capacitor (C2) is adjustable from the front panel. When the network has been properly tuned no adjustment is required over the 4-MHz range of the 2-meter band.

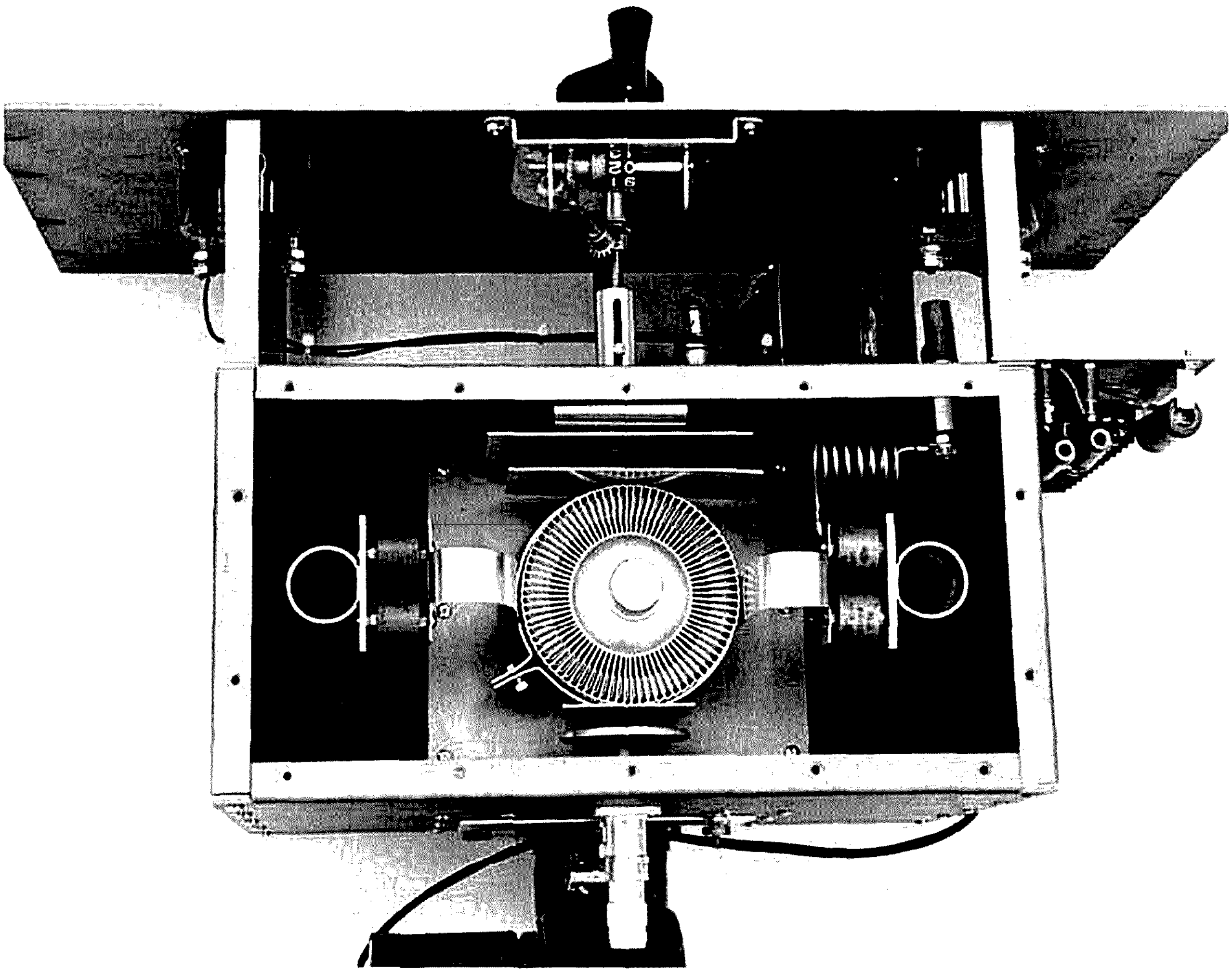
Underchassis layout of components is shown in the photograph. The cathode input circuit is in center compartment.

The slug-tuned coil in the input matching circuit is mounted on the rear wall. Air-wound filament chokes are placed in front of the socket. The cathode-heater rf choke is near the top edge of the enclosure. All of the cathode leads of the socket, plus one heater pin (pin 5) are connected in parallel and driven by the input matching network.

The ceramic socket for the 8877 is mounted one-half inch below chassis level by spacers. Four pieces of brass shim stock (or beryllium copper) are formed into grounding clips to make contact to the control grid ring. The clips are mounted between the spacers and the chassis. The aluminum clamps holding ends of plate lines are visible in the side compartments. The filament transformer and dial mechanism are placed in the area between inclosure and panel.

plate circuit

The plate circuit of the amplifier is a transmission-line type resonator. The line (L5 plus L6) is one half-wavelength long with the tube placed at the center (fig. 2). This type of tuned circuit has several advantages. A quarter-wave circuit would normally be preferred because of its greater bandwith, but I wanted to use easily obtainable standard copper water pipe as the center conductor of the transmissionline tank circuit. The resulting high-impedance transmission line would make a quarter-wave plate tank circuit physically short and difficult to handle.



Top view of amplifier showing plate compartment. 8877 tube is at center with plate lines on each side.

In addition, the heavy rf current that flows on the tube seals and control grid would, in the process of charging up the output capacitance to the plate voltage

this type of cavity is complex and difficult to build.

A practical compromise is to use two quarter-wave lines connecting to opposite

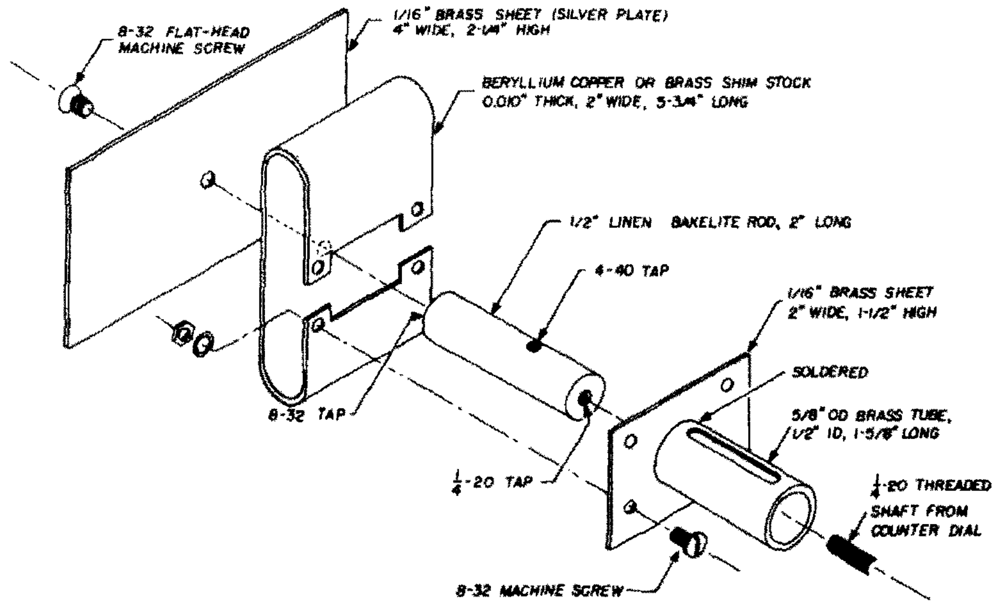


fig. 3. Variable plate portion of plate-tuning capacitor C3. This arrangement permits the capacitor to be adjusted under full power without "jumpy" tuning as there are no moving or sliding contacts which carry heavy rf current.

swing, tend to concentrate on one side of the tube if a single-ended quarter-wave circuit were used. This current concentra-

sides of the tube. It is interesting to note that each of the two quarter-wave lines is physically longer than if only one quar-

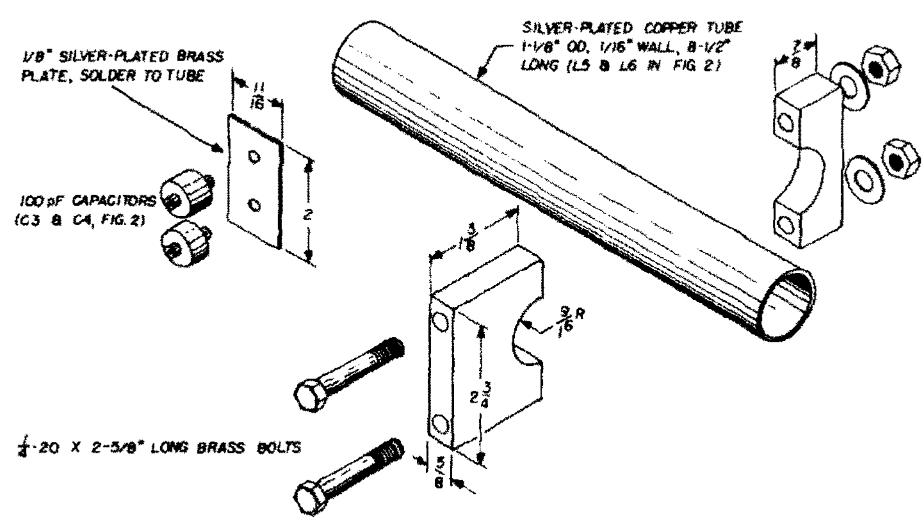


fig. 4. Details of plate lines L5 and L6. Copper tubes are standard copper water pipe.

tion would cause localized heating of the tube. The best tuned circuit configuration to minimize this effect is a symmetrical cylindrical coaxial cavity. Unfortunately,

ter-wave line were used. This is because only one-half of the tube output capacitance loads each of the two lines.

Resonance is established by a moving

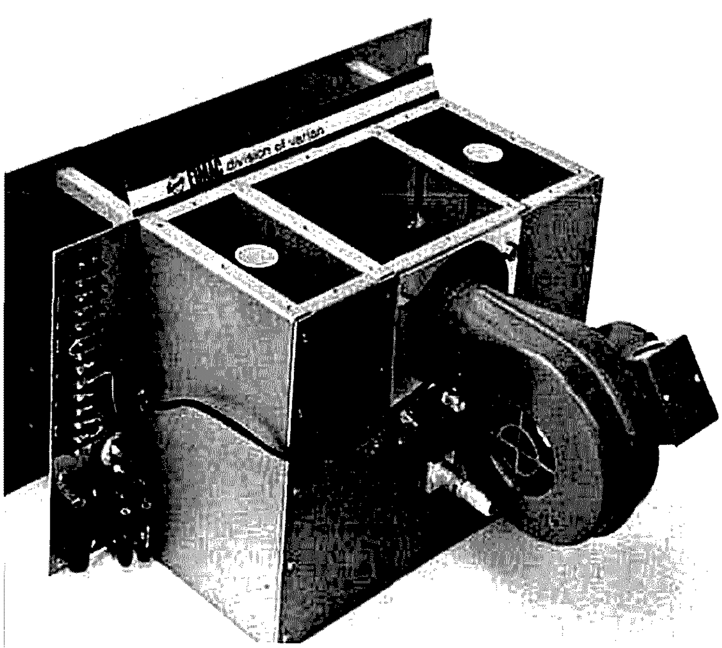
plate capacitor (C5); antenna loading is accomplished by a second capacitor (C6) placed at the anode of the 8877. Output power is coupled from the plate circuit through the series capacitor into a 50-ohm output. In the top-view photo tuning capacitor C_5 is at the front of the compartment; variable loading capacitor C_6 is at the rear. The plate choke is visible in the front corner.

construction

The two-meter amplifier is built in an enclosure measuring 10¼ x 12 x 6¼ inches. The 8877 socket is centered on a 6 x 6 subchassis plate. A centrifugal blower forces cooling air into the under-chassis area; the air escapes through the 2-5/8-inch diameter socket hole.

The plate tuning mechanism is shown in fig. 3. This simple apparatus will operate with any variable plate capacitor, providing a back and forth movement of about one inch. It is driven by a counter dial and provides a quick inexpensive and easy means of driving a vhf capacitor. The ground return path for the grounded capacitor plate is through a wide low-inductance beryllium-copper or brass shim stock which provides spring tension for the drive mechanism.

Rear of amplifier showing blower and coaxial output connector. Amplifier is upside down in this photograph.



The variable output coupling capacitor

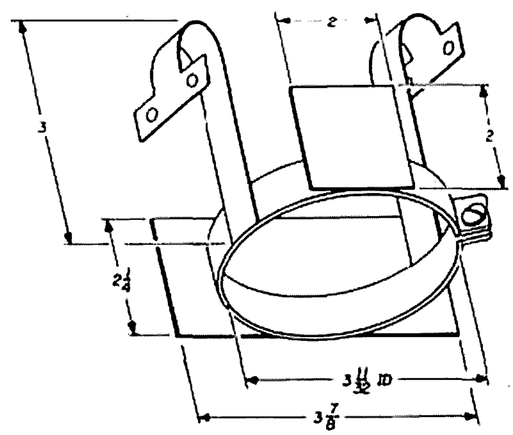


fig. 5. Anode clamp assembly for the two-meter linear amplifier.

is located at the side of the 8877 anode. The type-N coaxial fitting is connected to the moveable plate of the coupling capacitor. The fitting is centered in a special tubular assembly which allows the whole connector to slide in and out of the chassis, allowing the variable plate of the coupling capacitor to move with respect to the fixed plate mounted on the tube anode clamp. When the final loading adjustment has been set the sliding fitting is clamped by an arrangement similar to the slider on a variable wire-wound resistor.

The length of the plate-line inductors (L5 and L6) is adjusted by means of dural blocks placed at the shorted end of the line (fig. 4). The position of the blocks is determined by setting capacitor C5 at its lowest value and adjusting line lengths so that that plate circuit resonates at 148 MHz with the 8877 in the socket.

The plate rf choke is mounted between the junction of one plate strap and a pair of the dual blocking capacitors; the high-voltage feed-through capacitor is mounted to the front wall of the plate compartment. The blocking capacitors are rated for rf service, and inexpensive tv-type capacitors are not recommended for this amplifier. A short chimney to

direct cooling air from the socket through the anode of the 8877 is made from Teflon and clamped between the chassis deck and the anode strap.*

operation

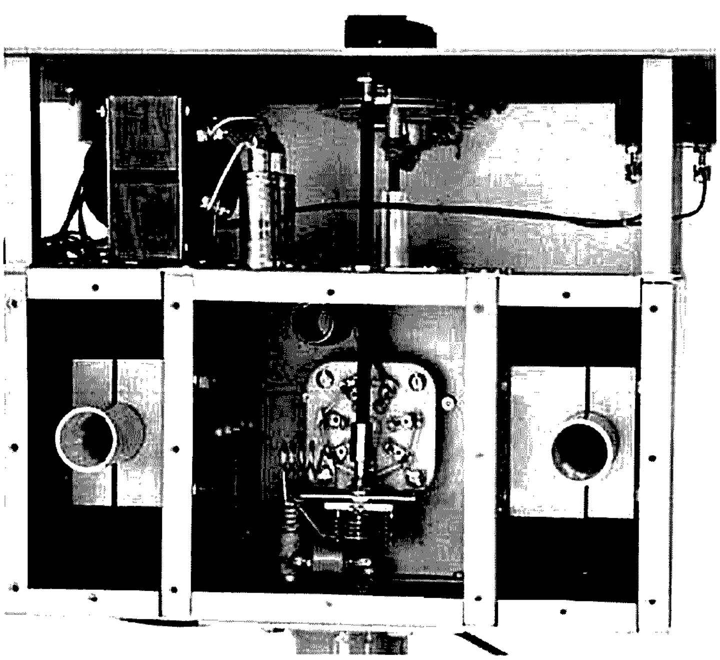
Amplifier operation is completely stable with no parasitics. The unit tunes up exactly as if it were on the "dc bands." As with all grounded-grid amplifiers excitation should never be applied when plate voltage is removed from the amplifier.

The first step is to grid-dip the input and output circuits to near-resonance with the 8877 in the socket. A swr meter should be placed in series with the input line so the input network may be adjusted for lowest swr.

Tuning and loading follows the same sequence as any standard grounded-grid amplifier. Connect a swr indicator at the output and apply a small amount of rf drive. Quickly tune the plate circuit to resonance. The cathode circuit should now be resonated. The swr between the exciter and the amplifier will not necessarily be optimum. Final adjustment of the cathode circuit for minimum swr should be done at full power because the input impedance of a cathode-driven amplifier is a function of the plate current of the tube.

table 1. Performance data for the 144-MHz power amplifier under the conditions most suitable for amateur ssb (2000 watts PEP) and cw (1000 watts).

Plate voltage	3000 V	2500 V	2500 V
Plate current (single tone)	667 mA	800 mA	400 mA
Plate current (idling)	54 mA	44 mA	44 mA
Grid voltage	-12 V	-12 V	-12 V
Grid current (single tone)	46 mA	50 MA	28 mA
Power input	2000 W	2000 W	1000 W
Power output	1240 W	1230 W	680 W
Efficiency (apparent)	62 %	62 %	68 %
Drive power	47 W	67 W	19 W
Power gain	13.8 dB	12.6 dB	15.5 d



Underchassis view of the two-meter amplifier. The cathode input circuit is in the center compartment. Plate lines are visible in the side compartments.

Increase the rf drive in small increments along with output coupling until the desired power level is reached. By adjusting the drive and loading together it will be possible to attain the operating conditions given in the performance chart in **table 1**. Always tune for maximum plate efficiency: maximum output power for minimum input power. It is quite easy to load heavily and underdrive to get the desired power input but power output will be down if this is done.

I would like to thank K6DC for his help in adjusting and determining the operating conditions for this two-meter amplifier.

*Detailed drawings of the anode clamp, plate resonator and blocking capacitor assembly, and variable plate tuning capacitor (C5) are available from R. Sutherland, EIMAC Division of Varian, 301 Industrial Way, San Carlos, California 94070. Ask for drawing numbers 168658, 168648 and 168647.

references

1. R. Sutherland, W6UOV, "Two Kilowatt Linear Amplifier for Six Meters," *ham radio,* February, 1971, page 16.

2. R. Barber, R. Rinaudo, W. Orr, R. Sutherland, "Modern Circuit Design for VHF Transmitters," *CQ,* November, December, 1965.

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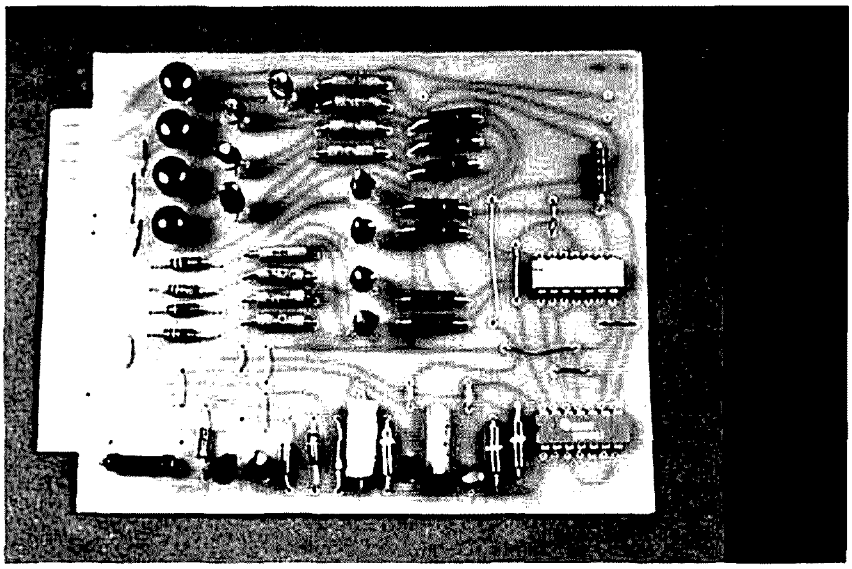


vhf fm channel scanner

With this low-cost solid-state fm channel scanner you can monitor up to four vhf fm channels at one time

George R. Allen, W2FPP, 4059 Bay Park Drive, Liverpool, New York 13088

Fixed-frequency amateur operation has become very popular during the last few years. One reason is the availability of commercial fm gear which can be converted easily to the vhf amateur bands. When using this equipment it's customary to pick a popular channel and monitor it for activity. This has the advantage that at any one time there will usually be a large number of local hams listening to the channel; to establish contact it is only necessary to give a short call.

While this type of fm operation may seem ideal it presents some problems. For example, a channel that is popular in one area may not be popular in another area. A mobile operator who is active on one channel has little chance of making contacts in another area where another channel is normally used. Furthermore, a channel which is popular one month may not be in vogue next month. Also, in a particular area you may find fm activity on a number of different channels. Without a large number of receivers, or a channel scanner, it's practically impossible to keep track of all the activity.

The four channel scanner described here can be used to monitor up to four channels in sequence. The scanner looks at each channel individually for 100 milliseconds. If there is no activity the scanner goes on to the next channel. This process continues until activity is encountered, at which time the scanner locks onto the active channel. It remains

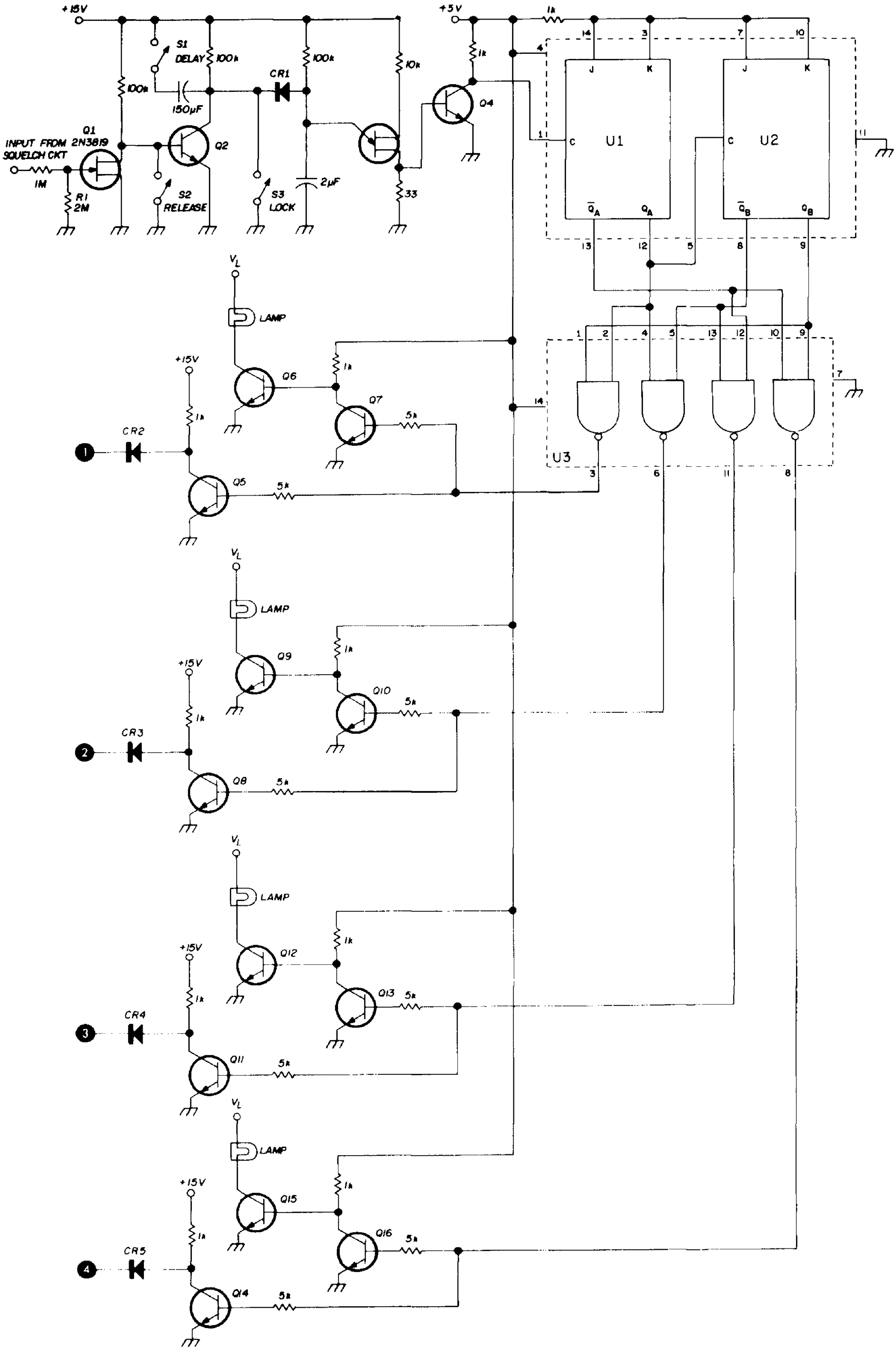

+15V
S1
DELAY
100k
CR1
100k
10k
+5V
1k
Q4
U1
U2
100k
150μF
Q1
2N3819
INPUT FROM SQUELCH CKT
1M
R1
2M
Q2
S2
RELEASE
S3
LOCK
2μF
33
LAMP
Q6
Q7
5k
CR2
Q5
U3
LAMP
Q9
Q10
CR3
Q8
LAMP
Q12
Q13
CR4
Q11
LAMP
Q15
Q16
CR5
Q14
VL

table 1. Parts list for CS-4 channel scanner.

qty	item	price	source
1	printed-circuit board	$3.50 (undrilled) 4.50 (drilled)	Alton Industries, 7471 Thunderbird Rd., Liverpool, New York 13088
1	SN7473 IC	1.00	Gateway Electronics, 6150 Delmar Blvd., St. Louis, Mo. 63112
1	SN7400 IC	.50	same as SN7473
1	unijunction transistor	.30	Radio Shack, 4 for $1.19
1	2N3819 fet	.50	Poly Paks, P. O. Box 942, South Lynnfield Mass. 01940, 2 for $1.00
10	npn small-signal silicon transistors	2.38	Radio Shack, 15 for $1.19
4	2N3641	1.19	Radio Shack, 4 for $1.19
5	small-signal silicon diodes	.02	Poly Paks, 40 for $1.00
24	resistors	2.40	Radio Shack, etc.
1	2-μF capacitor	.40	Radio Shack, etc.
1	150-μF capacitor	.60	Radio Shack, etc.
		$12.79	

locked on the active channel until manually released, or activity stops.

When designing the scanner I wanted to develop a reliable unit which could be built easily at low cost. These goals have been met; the scanner can be built for less than $13 if all parts, including the printed-circuit board, are purchased new. The unit can be completely assembled in 3 or 4 hours. In addition, the channel scanner, which I call the CS-4, can be connected to many of the fm transceivers on the market without making major modifications.

fig. 1. (Left) Vhf fm channel scanner. Lamps are 6, 12 or 24-volt, 200 mA maximum; voltage V_L is proper voltage for selected indicator lamps.

Q2,Q4,Q5,Q7, Q8,Q10,Q11, Q13,Q14,Q16	npn small-signal transistor (Radio Shack 276B590)
Q3	unijunction transistor (Radio Shack 276B111)
R1	optional 2-meg resistor (see text)
U1,U2	part of Texas Instruments SN7473 dual J-K flip-flop
U3	Texas Instruments SN7400 quad dual-input gate

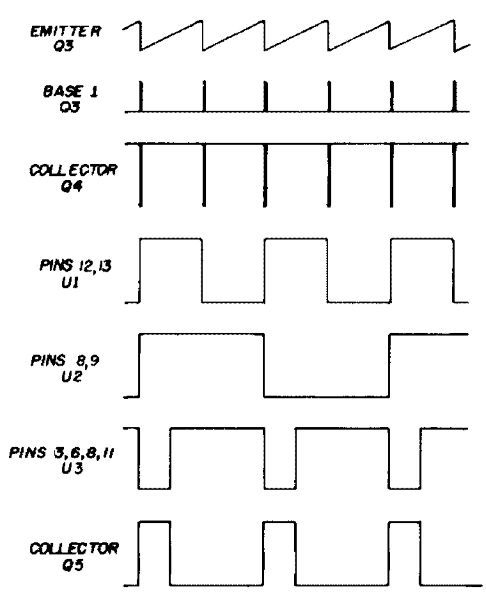


fig. 2. Typical waveforms of the channel scanner.

The CS-4 is built around two inexpensive integrated circuits: the Texas Instruments SN7473 dual JK flip-flop and SN7400 quad dual-input AND gate. The dual JK flip-flop counts to four using two the scanning rate is about 100 ms per channel. This rate seems adequate. However, if you want a different scanning rate the 2 μF capacitor may be changed; lowering the value will increase the scan-

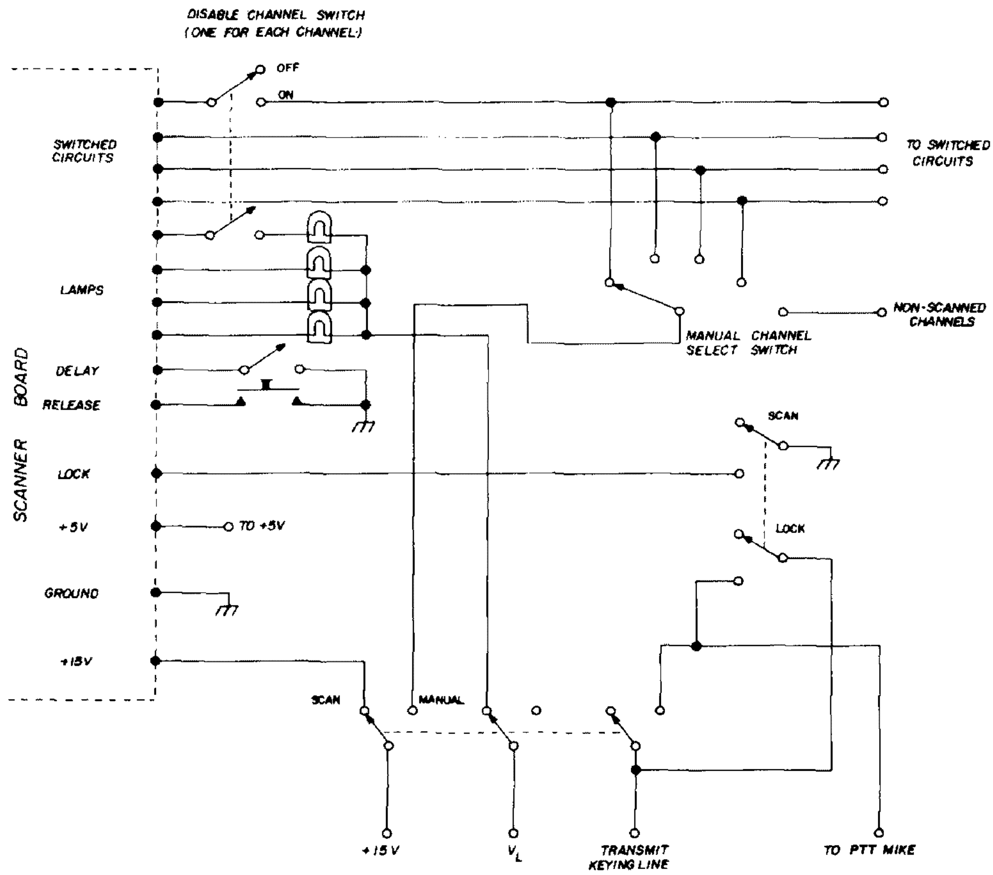


fig. 3. Channel scanner installation at W2FPP's station.

binary bits for the count. The quad dual-input gate separates the counts of the two binary bits into four separate outputs. The four separate outputs drive discrete transistors which in turn drive the lamps and the diode or transistor switches for each channel. Lamp drivers and diode or oscillator drivers must be used since the SN7400 will not handle the required voltage or current. Although an additional IC could have been used, one designed for output driving, discrete components were chosen to keep cost to a minimum.

Since the scanner is built around a counting circuit it's necessary to provide a source of pulses to be counted. A variety of circuits were tried, and most of them worked well; the unijunction circuit as shown in fig. 1 proved to be the least expensive. With the components specified

ning rate. If the scanning rate is too fast, however, the scanner will not lock since signal detection in the receiver may not occur rapidly enough.

The output of the unijunction clock circuit is fed into an npn transistor operating in the saturated mode. This transistor produces a large-amplitude negative-going pulse. The large negative-going pulse drives the flip-flops.

The channel-locking circuit was the result of considerable experimentation. Several problems were encountered: first was the problem of stopping the oscillator quickly; the second was how to get the receiver to stop it. One of the first approaches used an npn transistor with the emitter grounded and the collector connected directly to the emitter of the unijunction. A positive voltage caused the transistor to conduct, shorting the

emitter of the unijunction to ground, stopping the oscillator. However, this circuit had too much lag.

With scanning rates less than one channel per second the lock circuit did

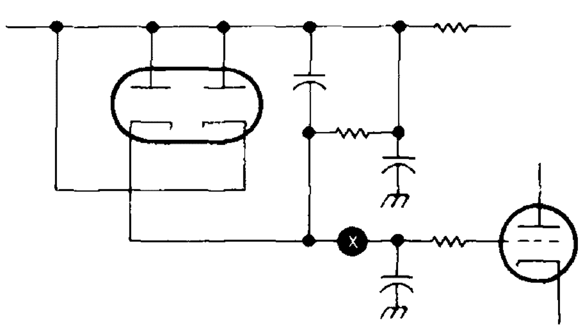


fig. 4. In a typical tube-type squelch circuit the channel scanner is connected at point X.

not react fast enough, so the scanner never locked. Although one channel per second may sound like a reasonable rate, it resulted in parts of a transmission being missed before lock occurred. This problem was corrected by using a small silicon diode, CR1, to isolate the emitter of the unijunction from Q2. This transistor switches the unijunction emitter from ground to +15 volts.

Solving the problem of locking the receiver turned out to be fairly simple. It was only necessary to find a point in the receiver that had a marked voltage change between signal and no-signal. Two points used successfully were the grid of the first limiter and the derived squelch-voltage output. It is preferable to use the squelch-voltage output since it is immune to noise. The first limiter voltage fluctuates with noise and will cause the scanner to lock in the presence of noise; this can be quite annoying.

Although the input locking circuit in **fig. 1** is designed to operate with a tube-type receiver that has a squelch voltage that ranges from slightly positive to several volts negative (with signal) the scanner can also be used with transistorized rigs that have positive squelch voltage or current.

*Printed-circuit boards are available from Alton Industries, 7471 Thunderbird Road, Liverpool, New York 13088. Drilled boards are $4.50; undrilled boards, $3.50. Completely wired and tested channel scanners are $33.50.

construction

Construction of the scanner is simplified if you use the commercially-available printed-circuit board.* If you use the components shown in the table the scanner can be built for less than $13. The npn transistors are not critical and virtually any small signal silicon types in your junk box will work. The integrated circuits are not critical either. The only criteria is that they be of the TTL variety, designed for a 5-volt power supply. If other ICs are used it will probably not be

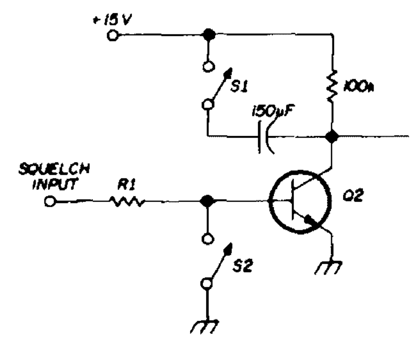


fig. 5. Channel scanner modification for use in solid-state fm receiver where signal voltage goes positive with signal. Q1, the 2N3819 fet input stage, is omitted. For value of isolation resistor R1 see text.

possible to use the commercial printed-circuit board since the pin connections will probably be different.

I recommend that you build one stage at a time and check it out before going further. With this step-by-step construction procedure it is easy to uncover possible problems. The unijunction oscillator stage should be built first. This circuit can be checked with an oscilloscope. The waveforms in **fig. 2** will help if you run into problems. If you don't have these waveforms check to see that all parts are wired correctly; then try parts substitution.

After the oscillator is working properly build the locking stages, Q1 and Q2. When a negative voltage of several volts is applied to the input of Q1 the oscillator should stop. Since the gate of Q1 will hold a charge for some time it may be

necessary to wire an external 2.2 meg resistor from the squelch input to ground when testing the unit on the bench. After the locking stages are working the remaining portions of the scanner can be built. The waveforms in fig. **2** should help in diagnosing any problems.

installation

The scanner installation at my station is shown in fig. **3**. This circuit permits either manual or automatic operation by merely throwing a switch.

The squelch input to the scanner can be obtained from a typical tube-type transceiver as shown in fig. **4**. If this is not possible the scanner squelch input should be connected to the grid of the first limiter stage. The 2-megohm resistor from the gate of Q1 is needed only if the scanner lock signal is taken from the first limiter.

If the fm receiver is a transistorized unit that does not have a voltage that goes from zero to at least 2 volts negative, the scanner must be modified slightly. Using a vtvm, find a spot in the receiver squelch circuit which shows a marked voltage change with the presence of a signal. If the voltage goes from zero to positive from no signal to signal, apply this voltage directly to the base of Q2 through an isolation resistor (**fig. 5**).

If the voltage goes from positive to zero from no signal to signal, then the input to the scanner should be modified as shown in **fig. 6**. In this case, Q1 is replaced by a junkbox npn silicon bipolar. The isolation resistor R1 is determined experimentally by selecting a value which will lock the scanner, but *not* affect receiver operation.

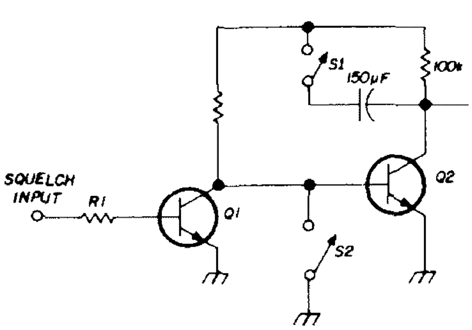


fig. 6. Channel scanner modification for use with solid-state fm receiver where signal voltage goes from positive to zero with signal. The fet input stage is replaced with an npn small-signal transistor. For value of isolation resistor R1 see text.

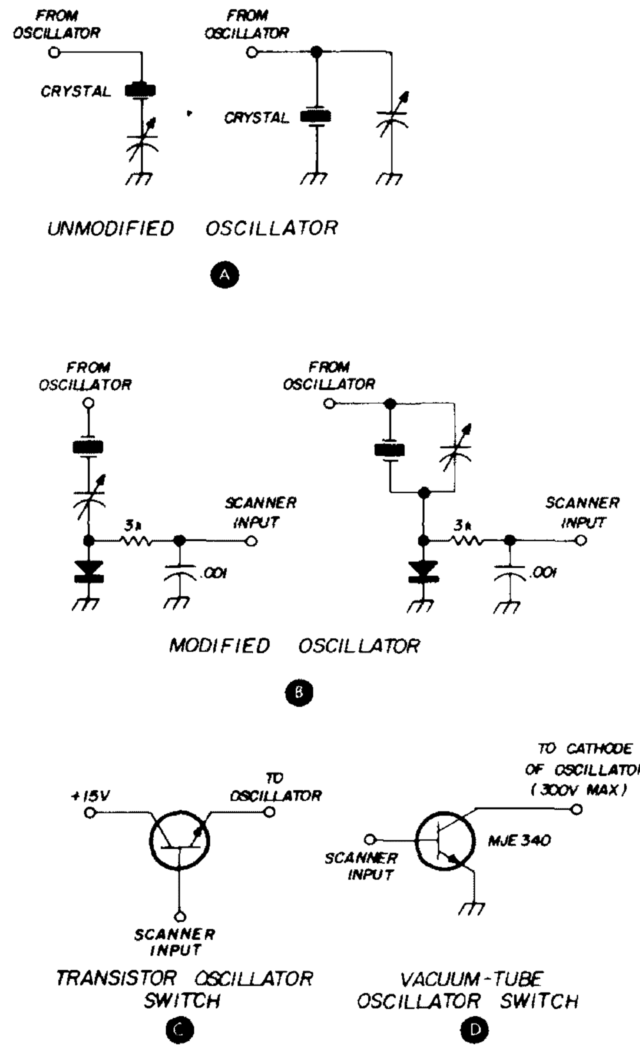


fig. 7. Methods of connecting the channel scanner to the receiver local oscillator stages. Typical modified and unmodified oscillator stages are shown in (A) and (B). Circuit for switching transistor oscillators is shown in (C); (D) is circuit for switching cathodes of vacuum-tube oscillator stages. Diodes in (B) are small-signal silicon.

To connect the scanner to the oscillator stages in the fm receiver the circuits must be modified slightly. One of the oscillator circuit modifications in fig. **7** should work with your receiver.

Two supply voltages are required for the scanner, +5 volts and +15 volts (nominal). The +5-volt supply is critical

and should be obtained from a regulated source. The nominal +15 volts may be anywhere from 12 to 18 volts.

using the scanner

To start scanning, all channel switches should be *on*, the scan-auto switch should be to *scan* and the delay switch to *out* . The channel scanner should begin scanning immediately. As soon as a signal appears on one of the four channels the scanner will lock. With the delay switch out, the scanner will start scanning again as soon as the signal on the channel disappears. In this mode you can keep track of activity on other channels during transmission breaks.

If you want to monitor both sides of a QSO turn the delay switch *on*. In this mode the scanner will remain on a given channel for several seconds after the activity ceases. This prevents switching and locking on another active channel during transmission breaks.

With the delay switch *out* the scanner will resume scanning for one scan if the release button is pressed when the scanner is locked to a channel. To transmit the scanner lock switch must be turned on. This locks the scanner on the active channel and the push-to-talk line is connected to the mike. You can get the same results by switching to *manual.* The manual position is used so that more than four channels may be used. At my station the transmit and receive oscillators for a given channel are tied to a single control line. When the scanner locks I am ready to transmit after setting the lock switch. By running the push-to-talk line through this switch the transmitter cannot be activated while scanning. If you operate the scanner while transmitting you'll get 100-ms pulsed signals on four channels.

summary

The CS-4 channel scanner was designed to provide a simple, low-cost fm scanner that could be installed with little difficulty. Many of these units are now in operation in central New York State with excellent results.

ham radio

vhf coaxial filter

A dual-stub system providing 32-dB attenuation at the design center frequency and more than 25-dB attenuation for a 1% frequency change

William I. Orr, W6SAI, Eimac Division of Varian, San Carlos, California 94070

More often than not, a good vhf receiving system is determined not by what it receives, but by what it rejects. The congestion of radar, radio positioning, and mobile services near amateur vhf bands often leads to distressing image problems. Many low-noise, transistorized receiving systems can be rendered near-useless because of a strong image signal from a station far removed from the amateur band.

On the other hand, spurious signals from improperly adjusted or poorly designed amateur transmitters can, and do, cause interference to other vhf services. Unwanted signals are everybody's problem. Such signals in the coaxial transmission line can be attenuated more than 30 dB by using dual stubs, one for unwanted signal rejection and the other to correct line impedance. These stubs are well worth trying, as they are inexpensive and they work! When interference is confined to a discrete range of frequencies, this high-Q coaxial filter will do an outstanding job of interference rejection.

the dual coaxial filter

The dual coaxial filter is shown in **fig. 1.** It works well between 20 – 250 MHz and has often been used in commercial mobile and point-to-point equipments.[1,2] Two stubs are used. One, l_1, is for signal rejection at the unwanted frequency; and another, l_2, is for impedance correction at the operating frequency.

The stubs are made from lengths of coaxial line. For low power (under 100

watts) and for receiving systems, RG-58A/U may be used. Transmitting systems should use the heavier RG-8A/U line. These two lengths of coaxial line, l_1 and l_2, properly adjusted, provide good filtering as shown by the unwanted-signal rejection plot of **fig. 2.** For a frequency change from the design frequency of one percent (1.4 MHz at 144 MHz), the dual filter offers better than 25-dB rejection to the unwanted signal. Tuned "on the nose," filter rejection is somewhat better than 32 dB.

operation

The rejection stub, l_1, is an electrical half wavelength at the rejection frequency. The impedance-correction-stub length, l_2, may vary from near-zero to an approximate electrical half wavelength at the operating frequency, as discussed later. Both stubs are placed in parallel across the coaxial transmission line at one point (preferably near the equipment) and are shorted at the far ends.

Since the rejection stub is one-half wavelength at the rejection frequency, its impedance across the coaxial line at this frequency is close to zero, effectively bypassing this frequency to ground. However, at the operating frequency, the rejection stub is not a shorted half-wavelength section and therefore presents either capacitive or inductive reactance to the transmission line. The unwanted reactance is compensated by adding the impedance-correction stub, which presents an equal and opposite reactance at the operating frequency.

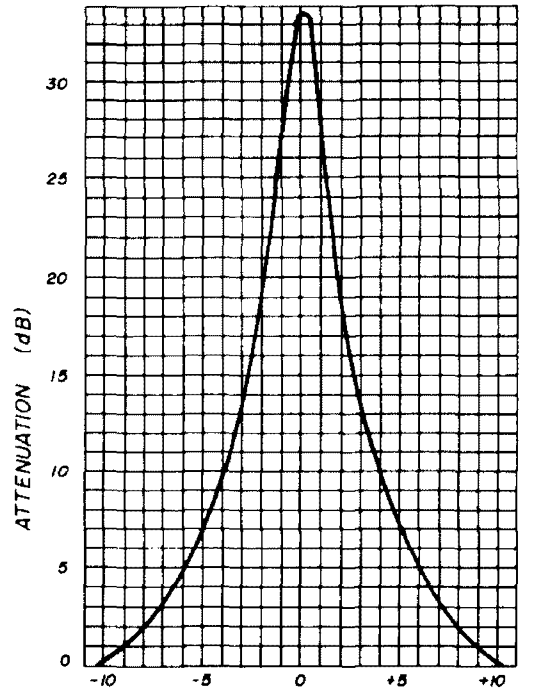


fig. 2. Attenuation characteristic of the dual coaxial stub filter. A single unwanted frequency may be attenuated as much as 33 dB by this simple filter, which offers over 25 dB attenuation at a frequency one percent removed from center frequency.

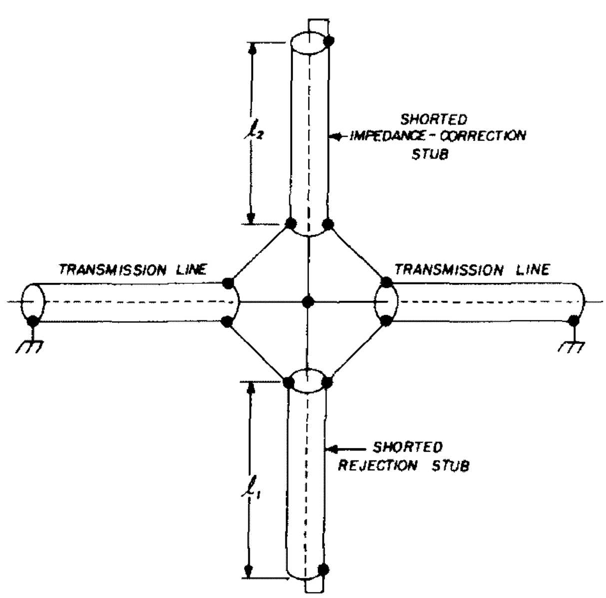


fig. 1. Dual coaxial stub filter made of shorted segments of transmission line shunted across regular transmission line. Presence of rejection stub is neutralized by impedance-correction stub.

The length of both stubs depends upon the operating frequency and the relationship between this frequency and the rejection frequency, as illustrated in **fig. 3.** For example, assume the operating frequency, f_o, is 146 MHz and the rejection frequency, f_r, is 118 MHz, a not uncommon relationship. The ratio 118/146 is 0.81f. This falls on the x-axis of the lower chart between 0.75f and 1f. The curve shows that, in this example, the reactance presented by the rejection stub to the line is inductive; and the upper curve shows that the impedance-

correction stub should present a capacitive reactance at the operating frequency, thus balancing the effect of the rejection stub at the operating frequency.

adjustment

Shown in **fig. 4** are approximate lengths for the stubs, as computed for solid dielectric cable having a velocity factor of 0.66. For the example, the rejection stub, l_1, is an electrical half wavelength at the rejection frequency of 118 MHz, or a little less than 40 inches long. A length of line a few inches longer

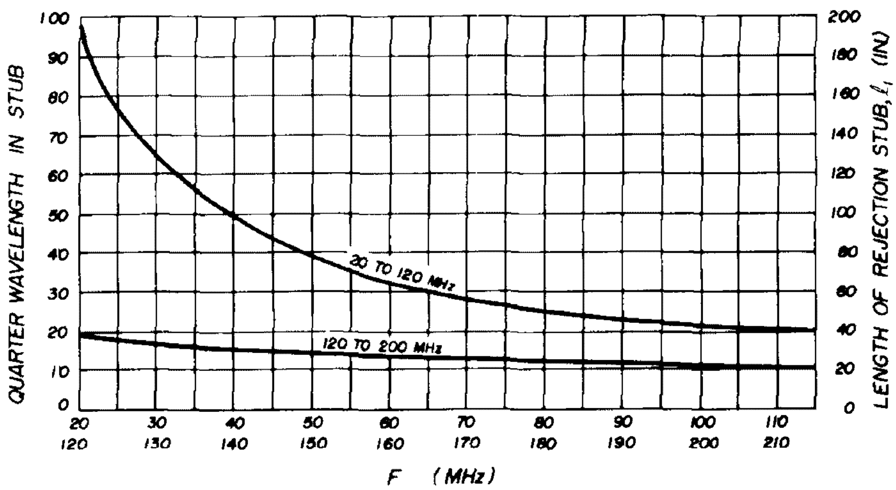


fig. 4. Correction and rejection stub lengths. Left-hand scale is used with fig. 3 to approximate the length of correction stub, since the stub is longer or shorter than a quarter wavelength at some frequencies. Right-hand scale is used for rejection-stub length.

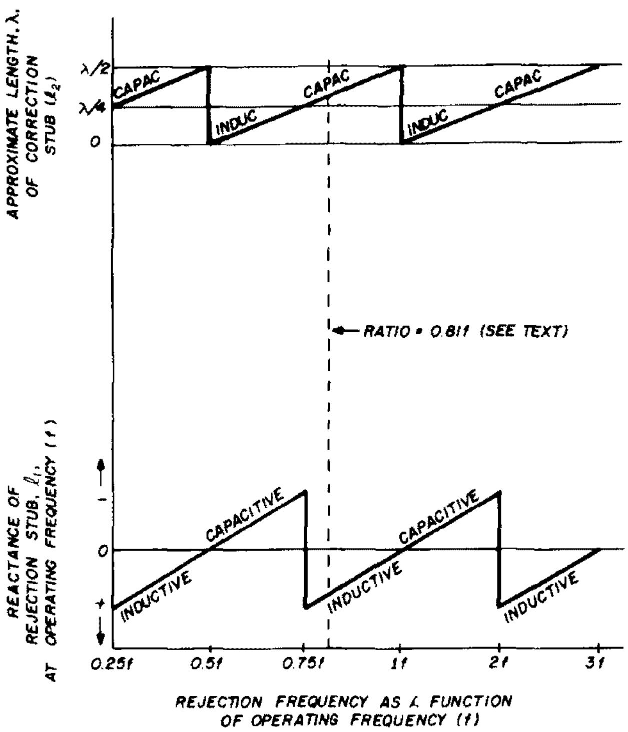


fig. 3. Reactance of rejection and correction filters as the ratio of rejection-to-operation frequency is varied. Net reactance of both filters at the operating frequency is zero. Sample filter for a rejection ratio of 0.81f is discussed in the text.

than this should be connected across the transmission line as shown in **fig. 1**. The stub may be adjusted to the rejection frequency with a grid-dip oscillator by placing a small loop at the unshorted end and trimming the shorted end for resonance.

Even easier, a signal at the unwanted frequency may be injected into the existing transmission line. Using a needle, the inner conductor of the stub is shorted to the outer conductor at various points, starting from the end of the stub and working toward the transmission line. The short is advanced about one-half inch at a time until the position of greatest attenuation of the unwanted signal is found. The line is permanently shorted at this point. This is best done by removing a short slug of the inner insulation and collapsing the outer shield around the center conductor and soldering it all around.

The impedance-correction stub, l_2, is now added. For various ratios of rejec-

tion-to-operating frequency, the length of this stub may vary from near-zero to about one-half wavelength. For an operating frequency of 146 MHz, a quarter wavelength in the coaxial stub is about 17 inches or less. If this stub were this length, it would present a very high

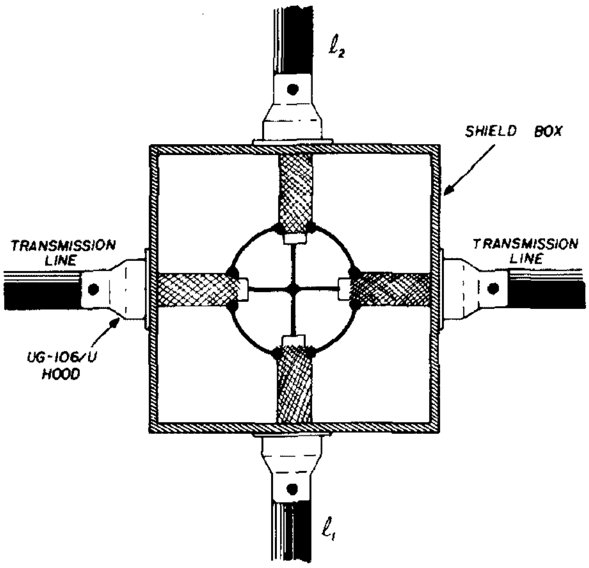


fig. 5. Typical assembly of vhf coaxial filter. Ends of coaxial lines are tinned and passed into brass box through UG-106/U hoods. Cable braid is soldered through holes in hood. The ends of the braid are jumpered together in the box with thin copper shim stock.

impedance to the desired signal. However, according to **fig. 3,** the correction stub should present a capacitive reactance and must therefore be longer than one-quarter wavelength. As a starter, the stub is cut to one-half wavelength and progressively shorted with a needle as outlined for the rejection stub. In this case, however, tuning is done to produce the least attenuation of the desired signal; or in the case of the transmitting stub, the least value of swr on the transmission line.

filter limitations

If the rejection frequency happens to be 1/2, 1/4, or 1/8 the operating frequency, the dual-stub filter cannot be used, as the rejection stub will be a full wave-length (or multiple thereof) at the operating frequency and will, therefore, present a near-short to the operating frequency as well as to the rejection frequency. On the other hand, if the undesired frequency happens to be twice or 3/4 the operating frequency, a correction stub will not be necessary since the rejection stub will be a quarter wavelength (or three-quarter wavelengths) at the operating frequency and will present a high impedance at this point.

construction

As any wise vhf enthusiast knows, rf has a disconcerting habit of running wild and free in the vhf spectrum. Unless properly contained, the unwanted signals can pass freely around the dual coaxial filter, rendering it useless. This means that no antenna currents should flow on the outer surface of the shield of the coaxial line and the shield should not be broken at the junction of the filters and line. The best way of attaching the two coaxial stubs to the transmission line, therefore, is to make up a small box of brass stock and mount a coaxial hood on the box for each cable, as shown in **fig. 5.** This makes an rf tight enclosure and helps keep the rf where it belongs.

references

1. Ramo and Whinnery, "Fields and Waves in Modern Radio," Section 10.11, Wiley & Sons, Inc., New York.
2. Craig, "A Coaxial Stub Filter," *Electronics,* June, 1951.

ham radio

Dear Zambo Electronic Parts Cleaner Co.— "I like Zambo cleaner because . . ."

easy-to-build integrated-circuit function generator

This easy-to-build function generator uses one IC to provide both square and triangular waveforms up to 1 MHz

Jim Fisk, W1DTY, editor

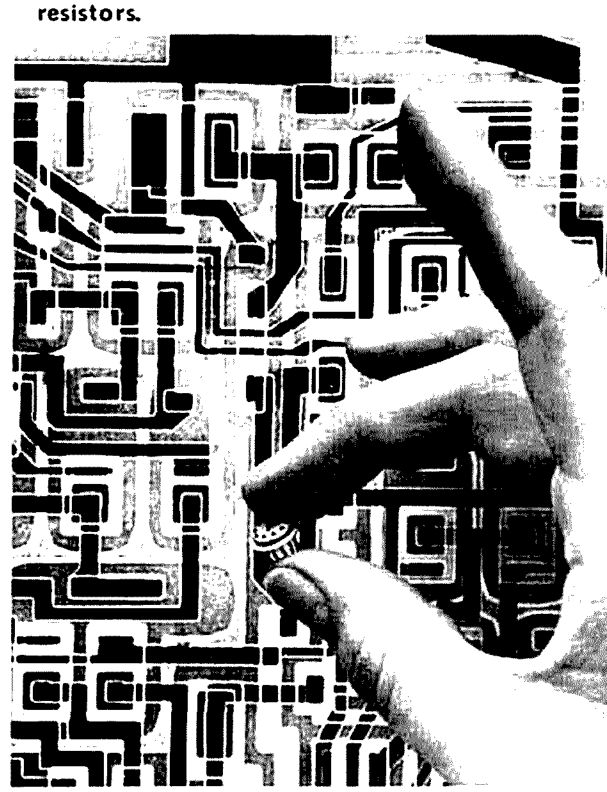
A function generator is a specialized signal generator normally used for testing complex electronic equipment; output signals include sine, cosine, square and triangular waveforms. As you might imagine, the electronic components necessary to generate this array of waveforms is rather complex. However, there is now an integrated circuit available that generates both square and triangular waves.

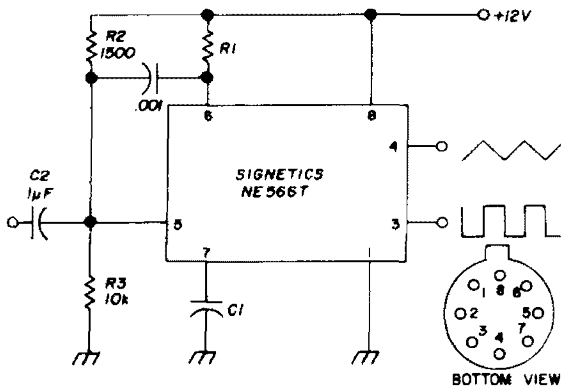
The new Signetics NE566 IC is a voltage-controlled oscillator that produces a highly accurate square wave and a very linear triangular wave up to 1 MHz. It can be used in tone generators, frequency-shift keyers, fm modulators, clock generators and signal generators. The output frequency is extremely stable.

frequency

The oscillation frequency is determined by the bias voltage on the control terminal (pin 5) and two external components, R1 and C1 (see fig. 1). The voltage at pin 5, as referenced to the positive

The Signetics NE566 function-generator IC contains 21 transistors, 10 diodes and 16 resistors.





C1 (uF)	frequency range (Hz)		
1	14	—	125
0.1	125	—	1250
0.01	1250	—	12500
0.001	12500	—	125k
100 pF	125k	—	1MHz

fig. 1. Basic function generator using the Signetics NE566T. Frequency is determined by R1, C1, and the bias voltage on pin 5.

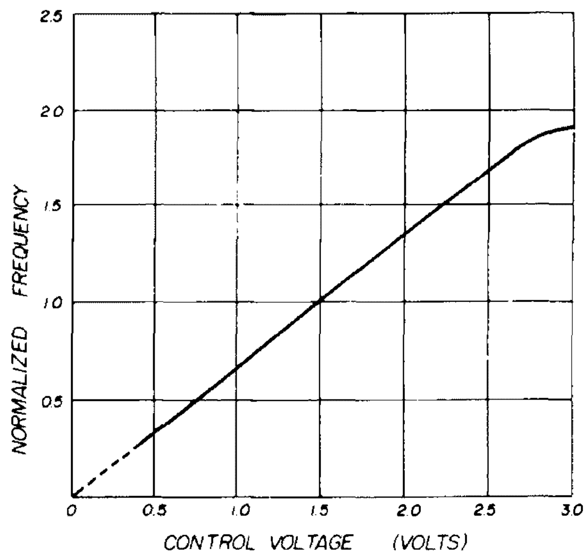


fig. 2. Normalized frequency output as a function of control voltage (control voltage is measured between pin 5 and pin 8).

voltage supply at pin 8, must be greater than 0.75 volt, but less than the positive supply (24 volts maximum). In **fig. 1** the control voltage is set by the voltage-divider resistors, R2 and R3.

Output frequency vs control voltage is plotted in **fig. 2.** *Normalized* frequency is used on the vertical scale to provide maximum usefulness. This means simply, that, if the oscillator frequency is 1000 Hz (with 1.5 volts bias), with 3 volts bias the output frequency will be approximately 1800 Hz. As you can see from this graph the integrated circuit can be modulated by the control voltage over a wide range with excellent linearity. A modu-

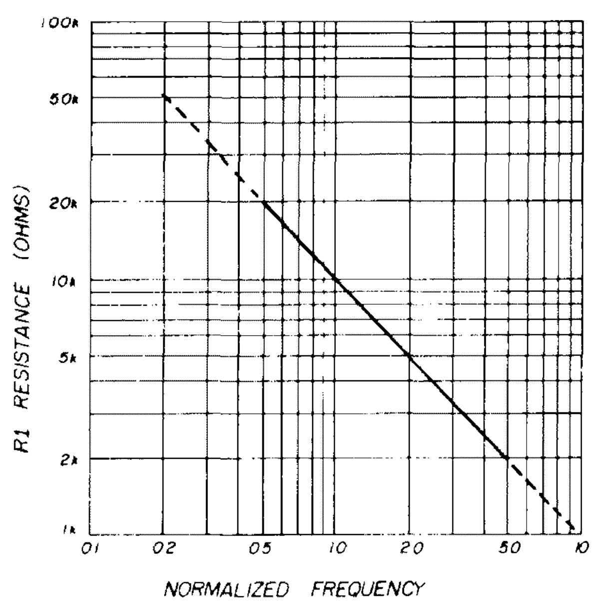


fig. 3. Normalized frequency output as a function of resistor R1 (with control voltage = 2 volts).

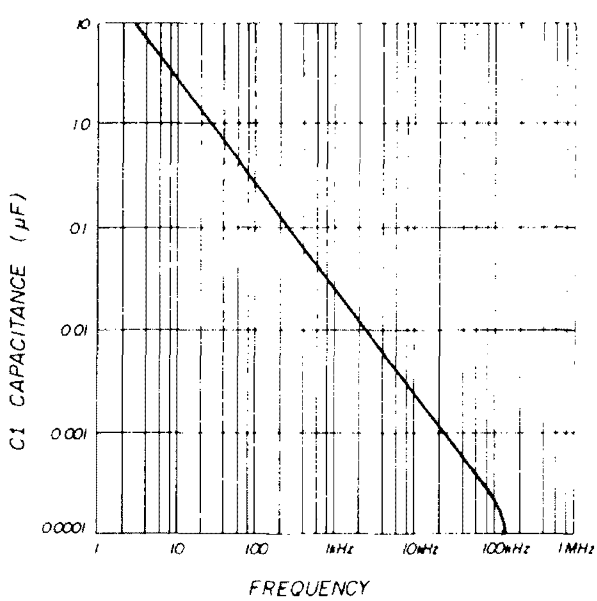


fig. 4. Frequency as a function of capacitor C1 (with R1 = 10k, control voltage = 2 volts).

lating signal can be ac coupled to the device through capacitor C2 in **fig. 1.**

Resistor R1 and capacitor C1 also determine output frequency. By proper selection of these two components the frequency may be set at any point up to 1 MHz. Once set, frequency drift is negligible at room temperature. Resistor R1 should be in the range between 2000 and 20k ohms. Its affect on output frequency is plotted in **fig. 3.** Note that *normalized* frequency is again used in this graph, with a normalized value of 1.0 when R1 = 10k. The value of C1 vs frequency is shown in **fig. 4;** this graph is based on 10k resistance at R1.

To show the use of **figs. 3** and **4** let's choose values for R1 and C1 that will provide an output of 3000 Hz. First go to **fig. 4** – 3000 Hz coincides with about 0.007 μF capacitance, a rather unordinary value. Choose the closest standard value, 0.01 μF. This moves the frequency down to about 2200 Hz.

If you read the fine print in the caption of **fig. 4** you'll see that this capacitor is based on a 10k resistance at R1. To bring the frequency up to 3000 Hz use a slightly lower resistance as shown in **fig. 3.** To find the necessary multiplication factor, divide 2200 into 3000; this gives a factor of 1.35. Entering **fig. 3** at 1.35 on the normalized frequency scale you find the required resistance value to set the output frequency at 3000 Hz; approximately 8000 ohms.

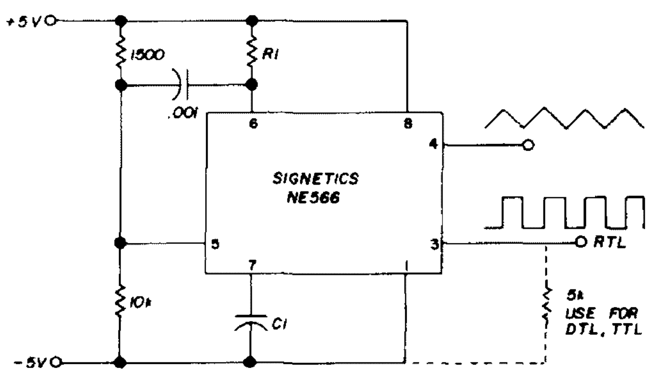


fig. 6. Plus and minus supply voltages put the square wave output at the proper dc level for driving logic circuitry. RTL logic may be driven directly; DTL and TTL logic require 5000-ohm resistor connected to the negative supply.

practical circuits

If the output frequency is off slightly when you build the circuit, you can add parallel trimming resistors to set the output frequency exactly. Or, if you want a variable frequency output, you can replace R1 with a potentiometer. A typical circuit is shown in **Fig. 5.** In this circuit capacitor C1 is selected by switch S1. Five capacitors will provide nearly complete coverage form about 20 Hz to 1 MHz.

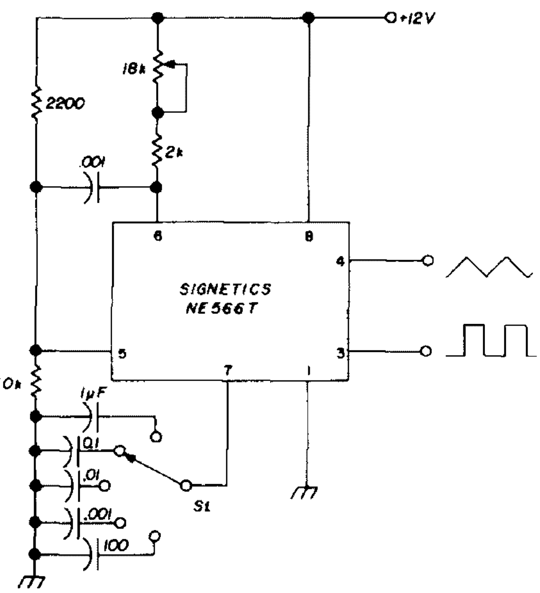


fig. 5. Variable-frequency function generator covers the frequency range up to 1 MHz.

When building this unit be sure to include the 0.001 capacitor between pins 5 and 6 of the integrated circuit (with *short* leads). This provents oscillation of the internal IC circuits.

The NE566T is a natural choice for driving standard IC logic circuitry. If you use the dual ±5-volt power supply as in **fig. 6** the square-wave output will have the proper dc level for driving directly RTL logic. For DTL or TTL logic connect a 5000-ohm resistor between pin 3 and the negative supply voltage.

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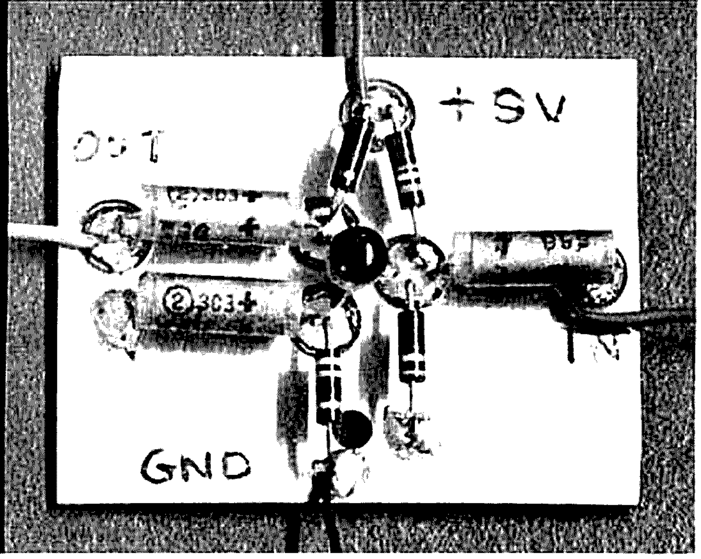


low-cost instant printed-circuit boards

Here's a simple technique for building printed-circuit boards for all your experimental projects

Ted Swift, W6CMQ, 1069 Savoy Street, San Diego, California 92107

The construction of electronic gear on a printed-circuit board presents a problem if a printed-circuit layout is not available. Breadboard assemblies for test prior to printed-circuit production, or one-time projects are examples of construction where it's hard to justify building a single printed circuit.

Instant tie points can solve most of these problems. By cutting a small ring through the copper surface of the copper-clad circuit board a small disc of copper remains which is firmly attached to the board but is electrically insulated from the rest of the sheet. These insulated discs make excellent tie points for connecting two or more component leads together. The tie points are very strong and, therefore, well suited for mounting components that can be supported with their own leads. Transistors, resistors, fixed or variable capacitors, inductors, crystal sockets, wire leads to external components and many other items can be soldered to these insulated discs. For example, three insulated discs, grouped close together, provide an excellent mounting base for a transistor and resistor, capacitor or other component. Terminals or pigtails that connect to ground are soldered to the copper surface of the circuit board *without* cutting a circular groove.

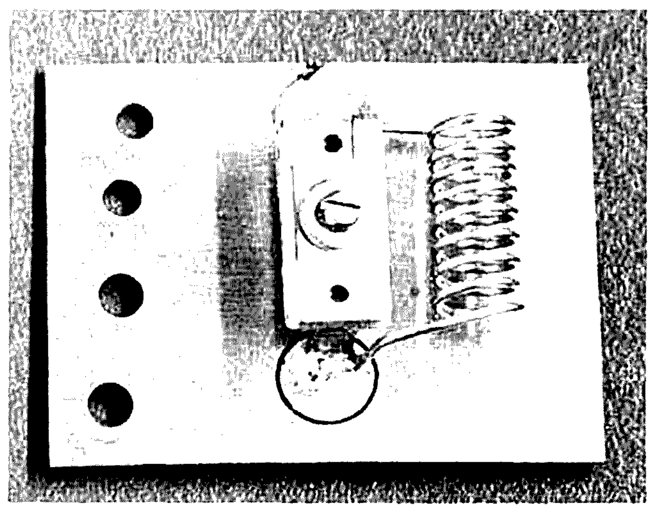
cutting tool

The best part of this construction technique is that the saw that cuts the ring in the copper surface can easily be made at home. Several different sizes, for producing large or small insulated discs, are handy. However, a description of the saw I use most often will serve as a guide for making other sizes.

A 2- or 3-inch length of mild steel rod, 1/4-inch in diameter, is the only material required. If necessary, true up one end of the rod by turning it in a lathe. If a lathe is not available, careful work with a hand file is an acceptable substitute. Bore a shallow hole in the end of the rod using a number 3 drill. This forms a cup in the end of the rod with a narrow rim. Now make two hacksaw cuts across the face of the cup, 90 degrees apart. File away the metal between adjacent saw cuts on a slant to form saw teeth; be sure the slant is in the direction that results in the cutting edge of each tooth being headed in the right direction when the tool is rotated in a drill press.

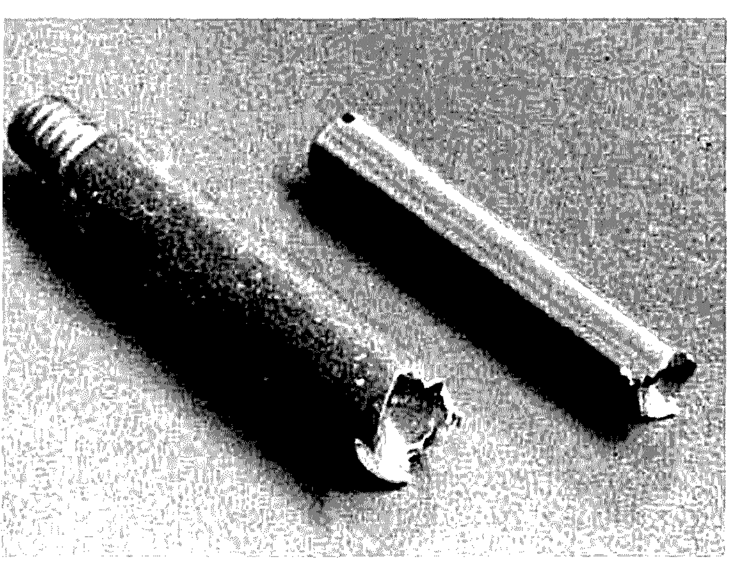
To use the cutting tool, chuck it in a drill press. Center the saw over the spot on the circuit board where the tie point is required. Cut through the copper surface and slightly into the insulating material of the board. Practice using the tool on scrap material to get the feel of it before attacking a project. If your saw is not too sharp it may leave a small burr on the copper surface. This may be smoothed out by rubbing the side of a round glass jar over the surface of the disc.

This tie-point technique is very flexible. Where practical it pays to plan the layout of all components on the circuit board in advance of construction. The required tie points can then be located and cut before parts are mounted, thus speeding the assembly job. However, if a tie point is forgotten, or if you need more of them, insulated discs can be added at any stage of construction without disturbing already installed parts.



Larger pads are useful for big components. This printed-circuit board is the same size as the transistor amplifier.

Insulated printed-circuit pads are cut with small saws made from mild steel rod.



Furthermore, the tie point technique has an unexpected advantage. Since the tie points can be located any place on the circuit board it is practical to mount components salvaged from surplus printed circuit boards. Such salvaged components usually have pigtails too short for reuse by other mounting methods.

One other circuit board hint: After the circuit board is cut to size and its copper surface is cleaned and polished with steel wool or household cleanser, spray the surface with a thin coat of *Krylon.* The copper surface will stay bright for life and solder will still wet the board as if the Krylon were not there.

ham radio

ac power-line monitor

Simple high performance circuit provides highly accurate line-voltage monitor that requires no calibration

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During the hot weather that persisted from last summer through early fall, the locality where I live experienced several power brownouts. Not blackouts, but a lowering of the nominal 117 volts ac in order to supply badly needed kilowatts to the adjoining metropolitan area. The measured line voltage dipped from its usual value of 120 volts to the vicinity of 110 to 112 volts. Air conditioners ran at reduced efficiency, lights dimmed somewhat and the automatic pin-spotter at the local bowling alley refused to function. Although some electrical appliances may not suffer from this sort of undervoltage it is a good idea to be able to read the line voltage with a higher level of accuracy than is possible with the average multimeter, which may be off as much as 6 volts (at 120 volts) and still be within tolerance.

There are several commercial ac voltmeters available for this purpose; most common is the iron-vane type. The better instruments are generally accurate to 2% of full scale. On a meter that reads 150 volts full scale this implies an inaccuracy of 3 volts at a line voltage of 120 volts ac. Since most of these instruments are not easy to read at a glance some commercial users and laboratories have changed over to the *expanded-scale* voltmeter. Simpson makes a fine segmental voltmeter with a mid-scale accuracy of 1/2%, but the high cost puts it beyond the reach of most amateurs and experimenters.

There are several electronic voltmeter circuits where the indicating meter reads all over-voltage above a certain standard. This standard or comparison voltage is a voltage-reference tube or zener diode. When properly calibrated the accuracy of such devices is in the neighborhood of 3%. There are two drawbacks, however: most voltage-reference sources are temperature dependent, and for reasonable accuracy the instrument should be calibrated with a laboratory standard.

The ac line monitor shown in fig. 1 neatly skirts these obstacles. It is simple and requires no calibration. In addition, the cost is lower than you might pay for an equivalent meter.

Basic to this ac voltmeter is the use of a segmental scale voltmeter that reads from 18 to 36 volts dc. My meter, a Marion Electric type HS2 was purchased

new at a surplus store for slightly under $4.00.* An ordinary silicon rectifier diode is used to furnish approximately 54 volts ac from the 120 volt line. Since the hermetically sealed Marion meter does not lend itself to easy modification I chose to use the 30-volt reading on the scale to correspond to 120 volts ac line voltage. The current drain at the 30-volt mark is 833 μA. (The 30-volt mark is 5/6 of full scale or 83.3%; since the meter has a 1-mA meter (the usual type) current flow at 5/6 full scale is 833 μA.

The value of multiplier resistor is determined with Ohm's law. Dropping 54 volts dc (from the rectifier) to 30 volts requires 28.8 kilohms for a current of

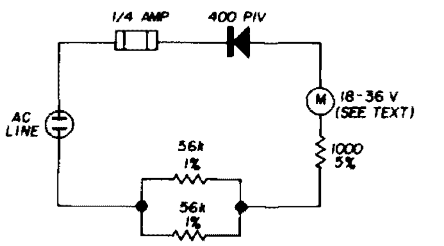

fig. 1. Circuit of the high accuracy ac line monitor.

833 μA. I used two parallel-connected 1% 56k resistors in series with an ordinary 1000-ohm resistor. The total resistance is very close to the desired value of 28.8k.

accuracy

Before talking about the performance of this simple instrument I'd like to explain some of the techniques used to insure that it will provide 1% accuracy without comparison to a lab standard.

First the dc meter was connected in series with four other 1,000-ohms-per-volt dc meters. One of these was a Weston 741 mirror-scale meter, rated at 1% accuracy. Readings were taken at 3-volt intervals; then an identical run was made backward from full scale deflection of 36 volts to 18 volts. Averages of the four other meters were tabulated; the readings of the expanded-scale Marion meter compared closely to the other instruments. The maximum error was 1.25% of full scale and the average error was less than 1%. After the instrument was completed ac accuracy was checked. A mirror-scale Simpson 261 vom was checked at 6.7 volts against a meter of known accuracy, a precision mirror-scale laboratory type ac voltmeter (Weston 433). Since the two readings were in agreement the Simpson 261 readings on the 0 to 10-volt ac scale were trustworthy. When switching to the 250-volt ac range additional precision multipliers are used; however, the basic ac circuitry and accuracy remain essentially the same. When my meter was compared to the Simpson 261, it read 119 volts as opposed to the Simpson's 119.5 volts.

You may wonder how such high accuracy may be obtained from such a simple circuit. Through segmentation, it is possible to increase the absolute accuracy of the original movement.[1] Proper selection of the multiplier resistors will contribute to the retention of this enhanced accuracy. As with any homemade device care and forethought are important ingredients that will pay off in increased accuracy and dependability.

The Marion meter movement is not highly damped. This allows a very slight flickering of the pointer but it is hardly noticeable. However, it has one definite advantage: Rapid excursions of the ac line voltage, not fully discernible with ordinary voltmeters, can be easily followed. I first noticed this effect when I hooked the ac line monitor to the wall socket where a small air conditioner was plugged in. When the compressor first started up, the ac line dipped – for an instant – to below 100 volts ac!

*The Marion Electric HS2, 2-1/2-inch round, 18- to 36-volt dc meter is available for $3.95 from Fair Radio Sales, Post Office Box 1105, Lima, Ohio 45802. A similar 18 to 36-volt Roller-Smith meter is available for $4 from Jeff-Tronics, 4252 Pearl Road, Cleveland, Ohio 44109.

reference

1. C. E. Miller, W1ISI, "Blow Up That Meter," *73*, October, 1964, page 56.

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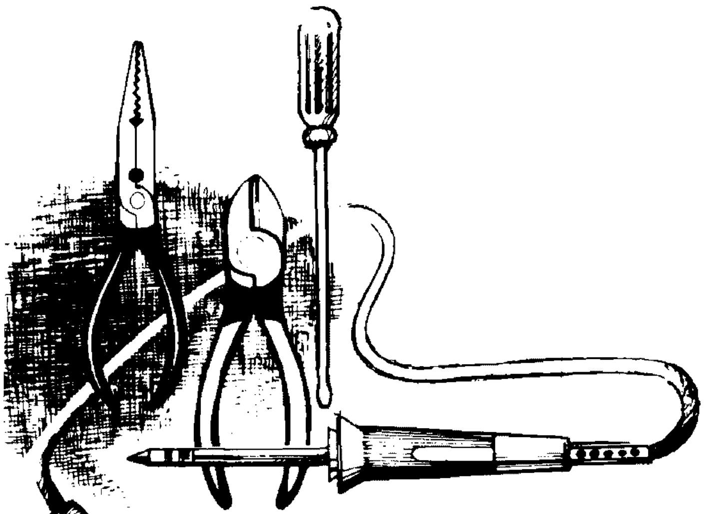


circuits and techniques

ed noll, W3FQJ

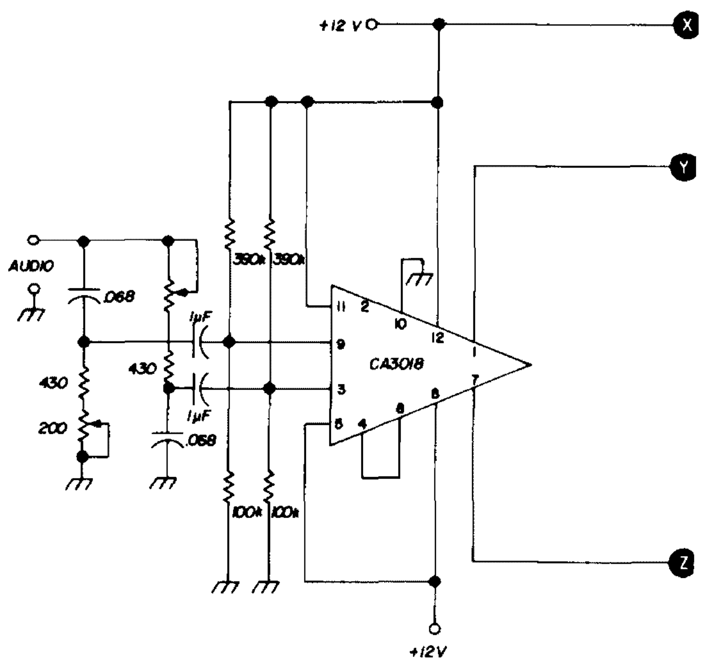
integrated circuits

Integrated circuits hold great promise for modulation and demodulation applications. The modern transmission modes of single sideband, double sideband, suppressed carrier and frequency modulation, plus phaselock receiving systems, depend a great deal on balanced circuits. This is the forte of integrated circuits with their arrays of identical diodes and identical transistors.

modulators

The diode balanced modulator is widely used in transmitters. IC diode-arrays provide the experimenter with small, well balanced diode assemblies. One example is the RCA CA3019, **fig. 1.** The CA3019 consists of four bridge-connected diodes plus two additional diodes, all in a miniature 10-pin TO-5 case. Conventional balanced modulator and ring modulator circuits are shown in **fig. 2.** In **fig. 2B** the diode between points **X** and **Y** consists of CA3019 diodes D1 and D2 in parallel; pin 5 is connected to **Y,** pins 6 and 2 are connected to **X.** Similarly, the diode between points **X** and **Z** consists of CA3019 diodes D3 and D4 in parallel; pin 8 is connected to **Z.**

Phasing-type single sideband generation may not disappear from the scene with the advent of integrated circuits. The balance complexities of the phasing system are reduced by the unusual uniformity that can be obtained in integrated-circuit construction. As shown in **fig. 3,** a combination of CA3018 and



CA3050 integrated circuits provide a packaged phase-shift single-sideband modulator.

The RCA CA3018 is a four-transistor IC that can be connected as two Darling-

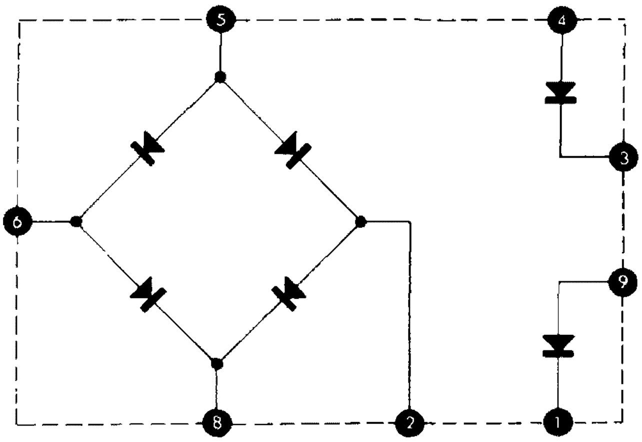


fig. 1. The RCA CA3019 diode array.

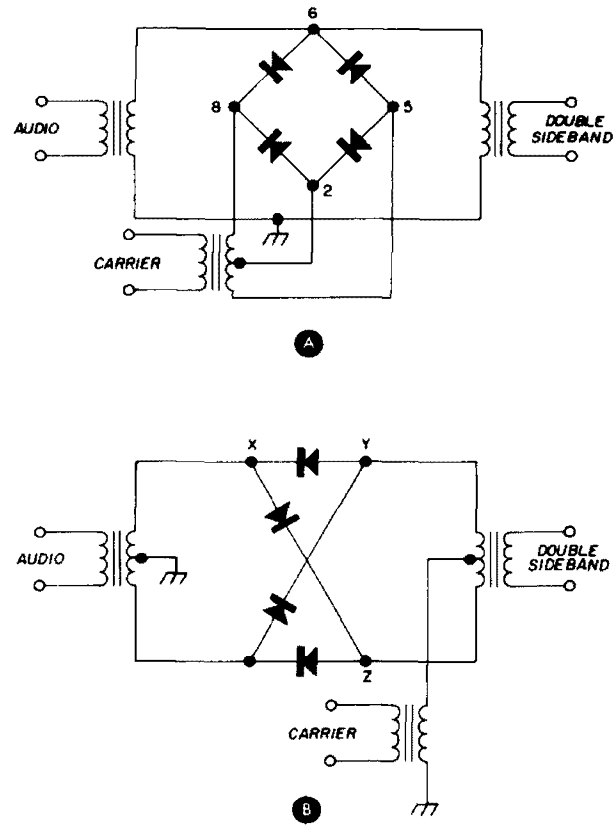
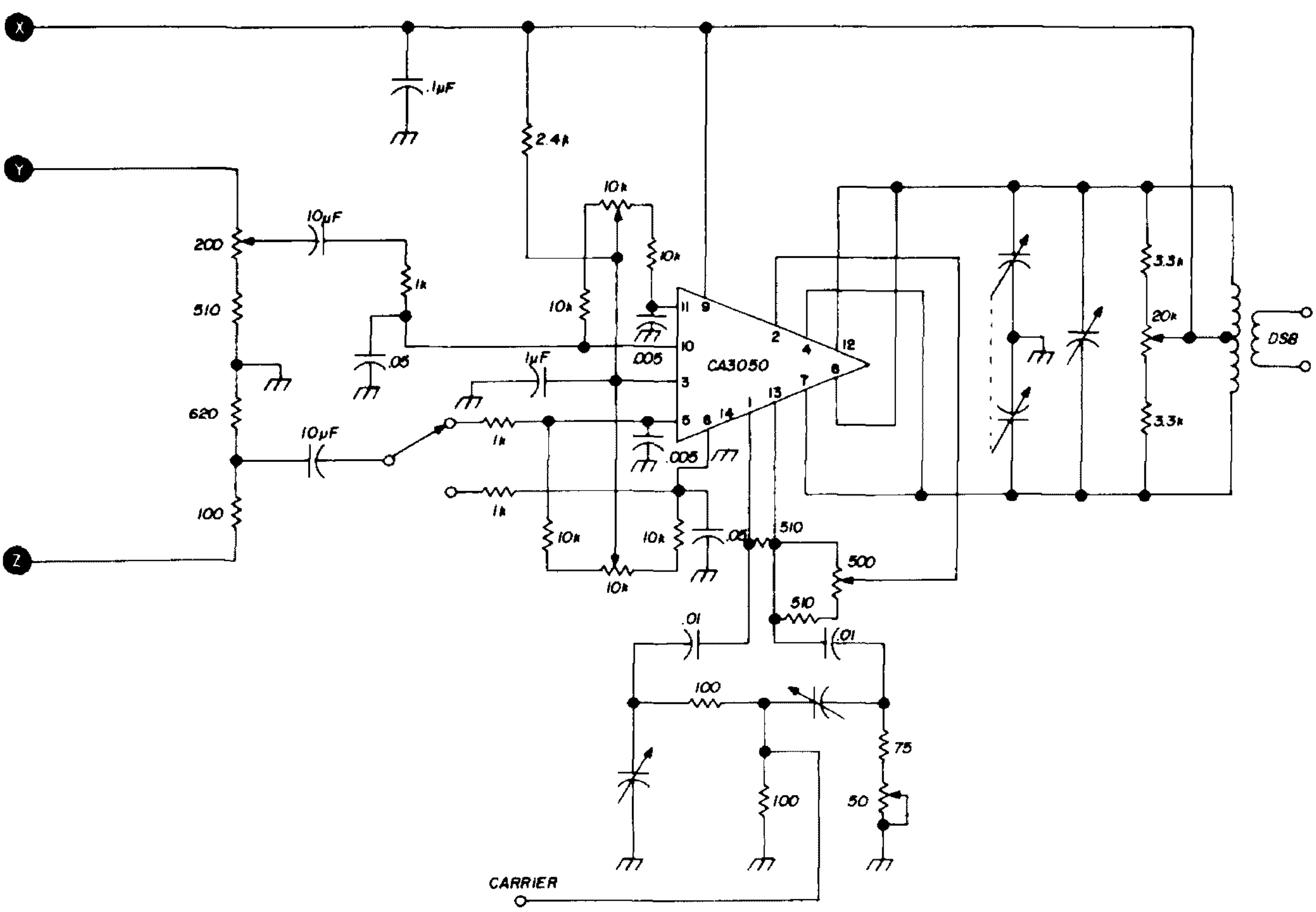


fig. 2. Balanced modulator (A) and ring demodulator (B) using the RCA CA3019.

ton pairs. The audio phase-shift network is connected externally, supplying 90°-related components to the separate inputs of the Darlington pairs; two low-impedance 90°-related outputs are made available. The Barker & Williamson 2Q4 is a good choice for the audio phase shift network.

The RCA CA3050 consists of two

fig. 3. Phasing-type ssb generator based on the RCA CA3018 and CA3050 integrated circuits.



independent differential amplifier arrays. Each array is made up of two Darlington pairs connected as differential amplifiers. Audio signals from the phase-shift IC are applied to their bases. A switch permits the selection of upper or lower sideband.

Additional transistors provide carrier input so the 90°-related carrier components can be applied to the emitter circuits of the independent dual differential amplifiers. The appropriate 90° phase shift networks are connected across the inputs of these transistors. One input consists of series resistor and shunt capacitor; the other, a series capacitor and shunt resistor. If each network provides 45° phase shift, the total phase shift of the two carrier components is 90°. The single-sideband signal is developed across the tuned output transformer.

demodulators

Some of the most intriguing ICs are those that contain an i-f amplifier plus capability for various demodulation modes. They lend themselves to the construction of small and versatile receiving systems.[1] With such IC versatility I wonder if a compact all-band (1.8 MHz to 144 MHz) all-mode, cw, a-m, fm and ssb receiver is a real possibility?

One particularly interesting IC is the National Semiconductor LM373 a-m/fm/ssb i-f strip. Simple external connections can switch the IC from one mode to another. Bandpass shaping may be accomplished externally over a frequency range from audio to 15 MHz. Agc is also included.

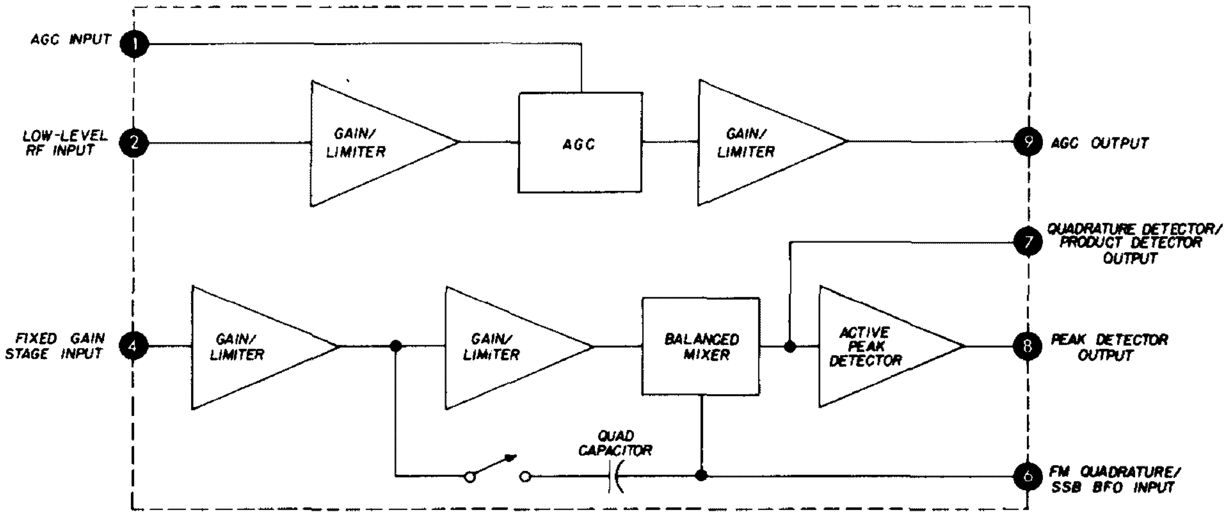


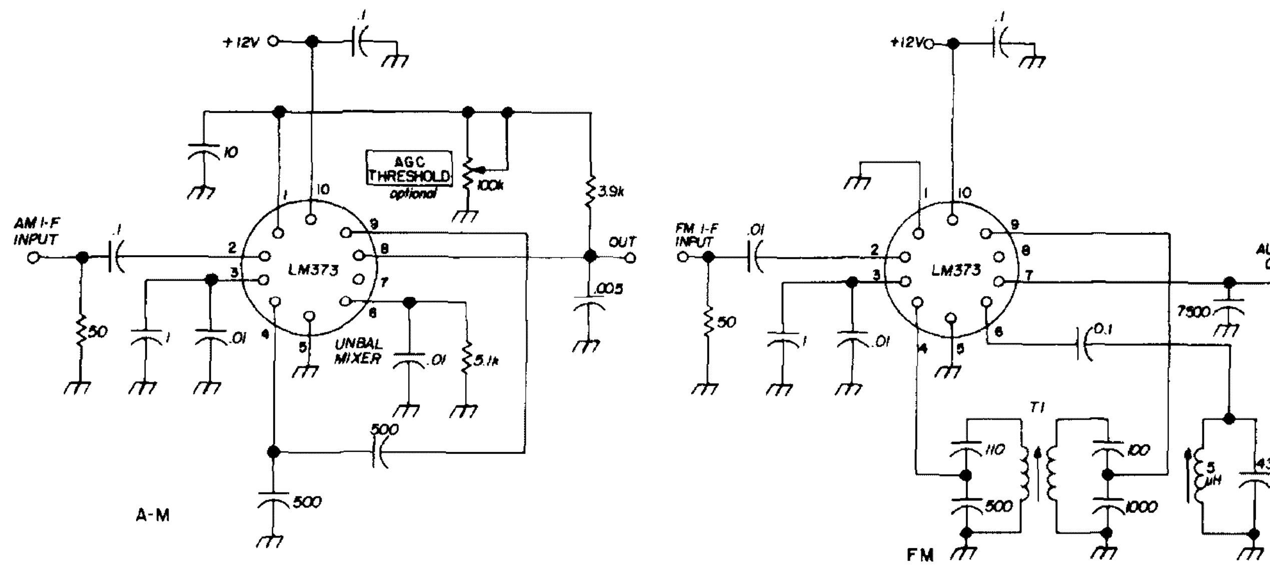
fig. 4. Block diagram of the National Semiconductor LM373 integrated-circuit a-m/fm/ssb i-f strip.

A functional block diagram of the LM373 is shown in **fig. 4**. The row of stages at the top include an amplifier/limiter, agc and second gain circuit. The signal to be demodulated is applied to pin 2. Output is taken at pin 9, and in most applications, re-introduced at pin 4, passing through an external filter circuit on the way. More gain stages are followed by a balanced mixer and peak detector. With proper external circuitry the latter two circuits can be switched for a-m, fm or ssb. Output voltage ranges from 50 and 120 millivolts rms depending upon mode.

Test circuit connections for a-m, fm and ssb are given in fig. 5. An a-m i-f signal is applied to pin 2, removed at pin 9 and re-introduced at pin 4. Demodulated output is taken from the peak detector at pin 8.

The fm detector uses a quadrature demodulator. The fm i-f signal is applied to pin 2, picked up at pin 9, and through transformer T1 is re-applied at pin 4. The quadrature LC circuit is connected at pin 6; demodulated audio is removed at the output of the balanced mixer at pin 7.

For ssb and cw operation the signal is sent through the amplifier/agc system, pin 2 to pin 9, and is re-inserted at pin 4. The balanced mixer is now used as a product detector with carrier applied to pin 6. Ssb output is again taken at pin 7. An appropriate switch permits selection of agc action for single sideband or manual gain control for CW.



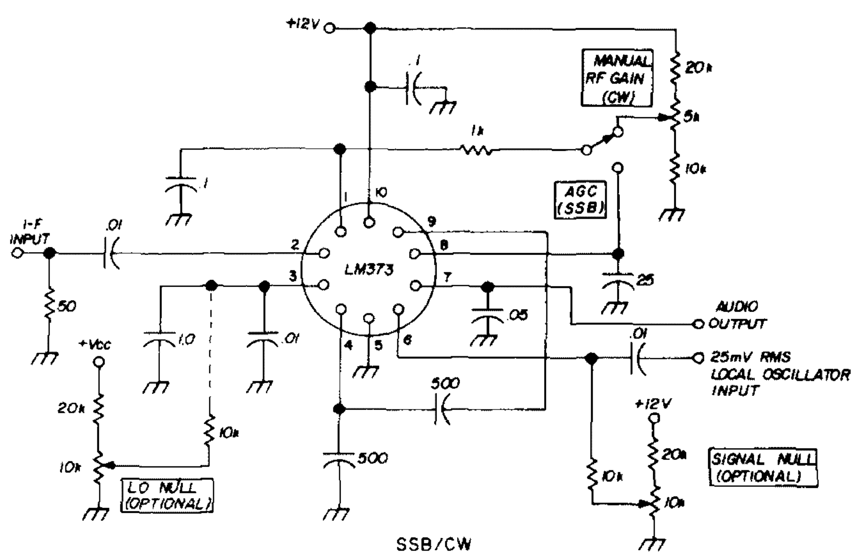


fig. 5. Circuits for using the LM373 integrated circuit for demodulating a-m, fm and ssb/cw.

rf amplifier

Integrated circuits are not the monsters that you may have been led to believe. That they destroy ingenuity and the yen for experimentation is more evasion than fact. Ingenuity is required to adapt these unusual devices for amateur applications and experimentation to make the ideas work. G3VJN points out in Technical Topics[2] that an amateur can often adapt some of these newer devices for applications very different from the manufacturer's intended use. To prove the point he built an rf linear amplifier for 80 meters using the popular RCA CA3020 *audio* power amplifier (fig. 6). A close check of CA3030A specs indicates operation to 8 MHz.

A diagram of the CA3020, fig. 7, shows the isolating input stage (Q1), differential amplifier (Q2 and Q3), driver (Q4 and Q5) and power output stage (Q6

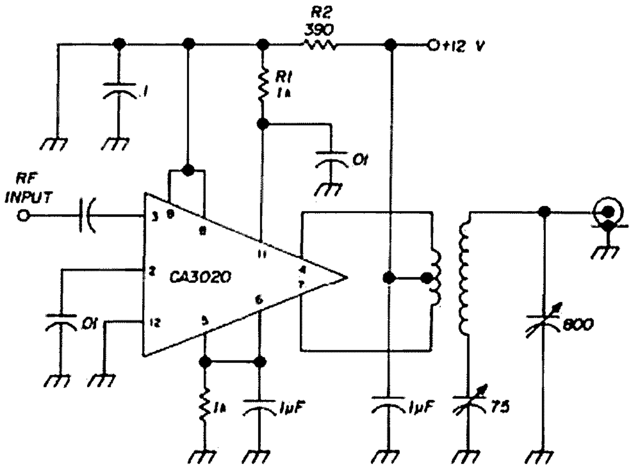
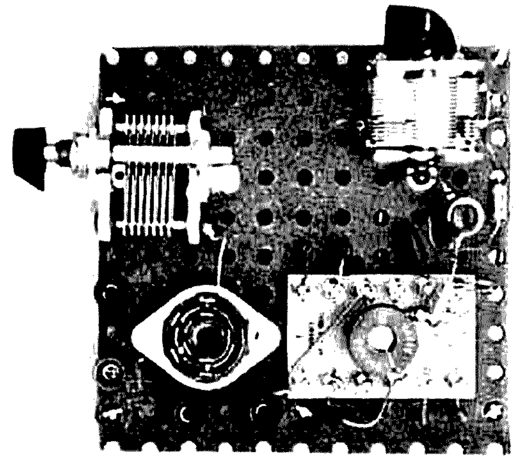


fig. 6. Integrated-circuit 80-meter linear amplifier designed by G3VJN.

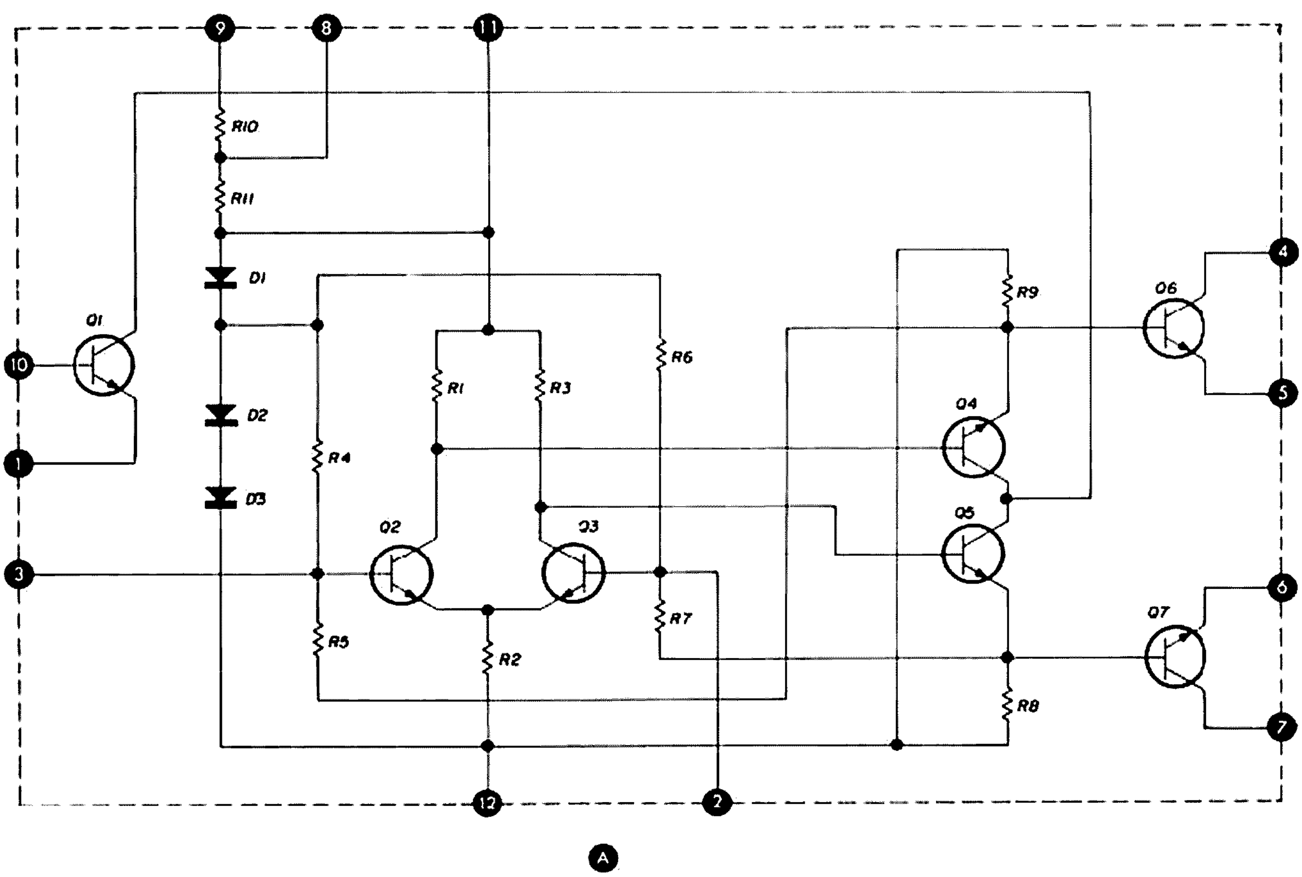


Breadboard layout of the RCA CA3020 linear amplifier.

and Q7). The IC also includes a built-in voltage regulator.

We can now look at G3VJN's circuit in fig. 6 and understand its operation. First of all, input transistor Q1 is not used; signal is applied to pin 3 which is the input to the differential amplifier. Supply voltage is applied to this stage through resistors R1 and R2 to terminal 11.

Push-pull output is taken from the final amplifier at terminals 4 and 7 which connect directly to the collectors of the output transistors. Supply voltage is applied through the coil center tap. The circuit is simple and easily understandable.



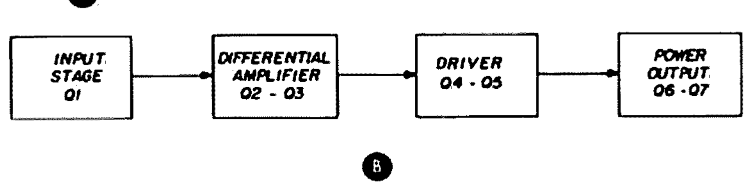


fig. 7. Block diagram and internal circuit of the RCA CA3020.

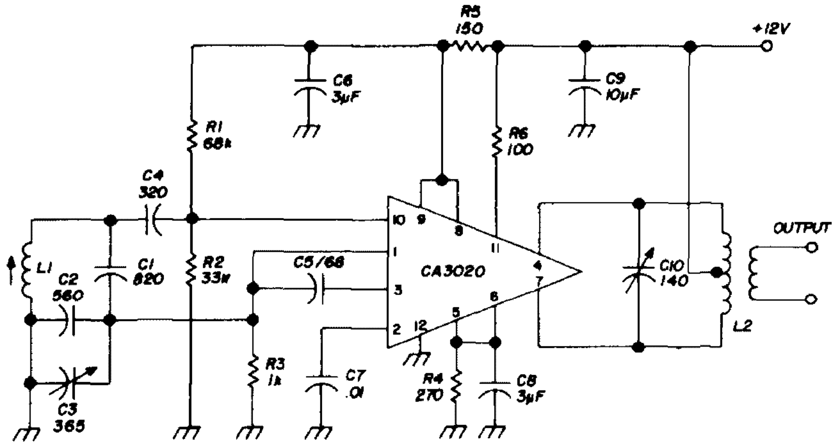


fig. 8. Seiler vfo and output amplifier using the RCA CA3020.

vfo

With the CA3020 you also have the makings of an integrated-circuit vfo, using the input transistor in a Seiler-type vfo. The circuit shown in fig. 8 performs well on 160, 80 and 40 meters. Since linear operation is not required for vfo operation G3VJN's basic circuit can be modified somewhat to obtain slightly more output.

The oscillator is built around pins 1 and 10 which connect to the base and emitter of the input transistor (fig. 7). The emitter output of the Seiler oscillator, pin 1, is connected to the input of the differential amplifier, pin 3, through capacitor C5. Some adjustment of C5 may be needed to obtain optimum output and minimum turn-on drift. Pins 4 and 7 connect to the push-pull rf output transformer. This is a center-tapped primary winding with secondary wound between the two halves of the primary.

A breadboard test unit of this circuit is shown in the photo. To facilitate circuit changes I used a Vector 570-F IC test socket. With this test socket the IC pins are brought out to 12 pressure spring terminals, a real blessing for experimental work.

The RCA CA3020 can also be used as an oscillator and frequency doubler to provide even greater isolation between the output and the oscillator. The circuit of fig. 8 frequency doubles by tuning the output circuit to twice the oscillator frequency.

Greater output can be obtained from the circuit of fig. 9. In this circuit the output transistors are connected in a push-push frequency doubler circuit. The bases are connected internally in a push-pull circuit that cannot be altered. The push-push doubler circuit requires that the collectors of the output transistors be connected in parallel as in fig. 9. A single-ended output circuit is used with

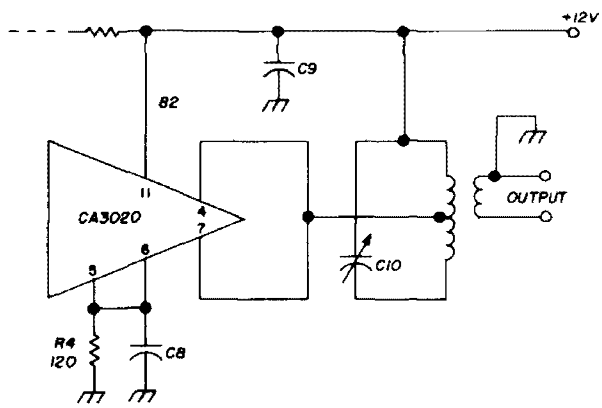


fig. 9. Modifying the output circuit of fig. 8 for frequency doubling.

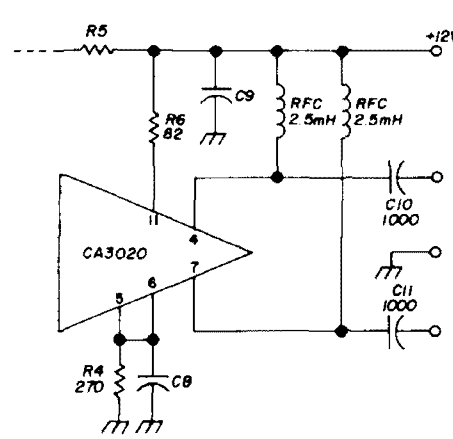


fig. 10. Modifications to the output circuit of fig. 9 for balanced voltage drive to succeeding high-impedance stages.

the collectors connected to the primary tap. Several hundred milliwatts rf output is possible.

The CA3020 circuit arrangement in fig. 10 is ideal for driving a succeeding high-impedance amplifier such as the push-pull input to a vacuum tube or fet. The primary objective of this circuit is to provide good voltage drive to the next stage.

triangle antennas

The full-wave triangle antenna seems like the best "hunk-of-wire" antenna ever; performance-per-penny-spent is exceptional. Only a single mast or other high support point is needed. Since it is fed at the bottom the triangle can be trimmed to resonance with ease.

The basic triangle is a closed full-wavelength antenna, fig. 11, with three equal-length sides. However, within reason, good performance is maintained if the base is longer than the two sides. Conversely, the sides can be longer than the base, within limits, as long as the overall length is one wavelength. The relative length of sides and base influence radiation resistance, therefore simplifying impedance-matching problems.

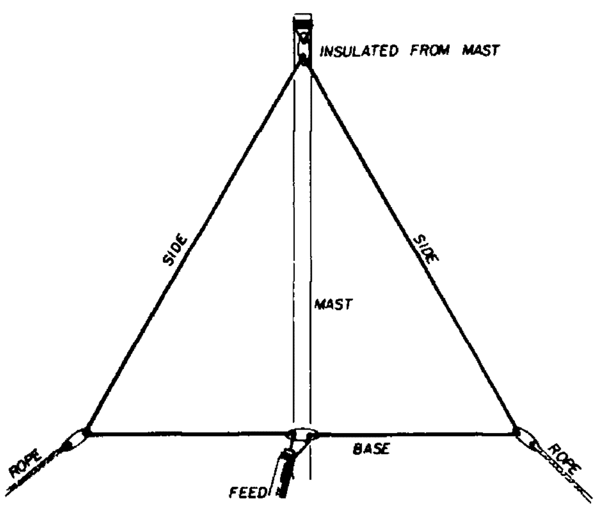


fig. 11. Basic triangle antenna layout for 40, 80 and 160 meters. Total length of wire is 1-wavelength on the band of interest.

On 40, 80 and 160 meters you can use coaxial feedline without any supplemental matching system. Simply make certain that the base of the triangle is 8 to 12 feet above ground; at this level the antenna is easy to trim for proper resonance and matching to a given transmission line.

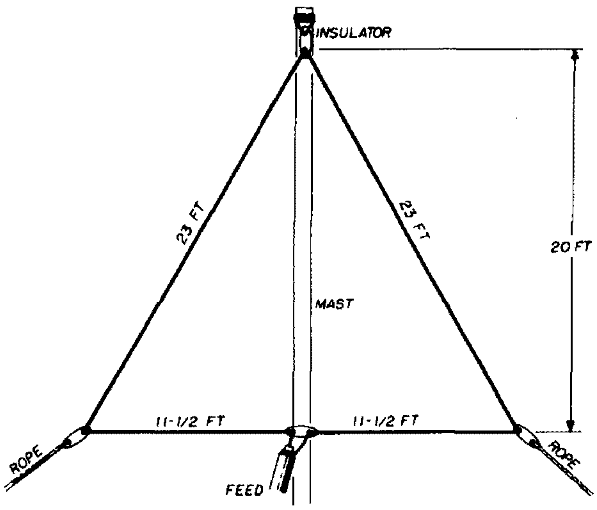


fig. 12. Triangle antenna for 10, 15 or 20 meters is mounted above ground on a single supporting mast.

construction

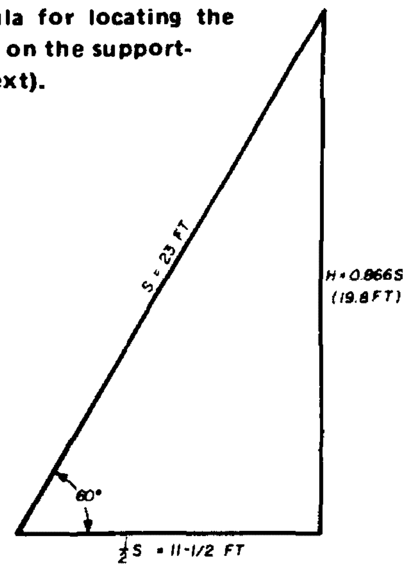
The mounting arrangement consists of a single support mast and a pair of ground stakes. The antenna wire is fed through an insulator at each end of the base and is supported by an insulator at the top of the mast. I use insulated wire such as plastic-covered no. 12, no. 14 or no. 16. The insulation is removed only where the antenna is connected to the transmission line at the center of the base. Nylon rope or plastic clothesline (non-metallic core) is attached to each of the base insulators; the triangle is pulled open and held in position by the two ground stakes.

On the higher bands, 10, 15, 20 meters, a single triangle antenna element can be supported in about the same manner, fig. 12. On these bands use coaxial line or open-wire transmission line and an antenna tuning unit. This provides good performance over the entire band for which the triangle is designed.

To build the high-band triangle measure off three equal sides. Attach the

two base insulators so they cannot slide along the antenna wire. Locate the exact position of the feed point. This can be done with the simple equation in **fig. 13.**

fig. 13. Formula for locating the base feedpoint on the supporting mast (see text).



For example, a full wavelength at 20 meters is:

$$\lambda = \frac{984}{f_o} = \frac{984}{14.2} = 69 \text{ feet}$$

Each side, S, will be one-third this value or 23 feet.

To set up an equilateral triangle, the separation between the apex and the center feedpoint of the base must be:

$$h = 0.866S = 0.866 \times 23 = 19.8 \text{ feet}$$

A twenty foot separation between apex and feedpoint is fine.

In erecting the antenna the apex and center feedpoint are attached to the mast. Then the mast is raised. Finally, the ropes attached to the base insulators are used to pull the wire into a triangle.

The triangle antenna has a bi-directional pattern and provides good performance at reasonably low mounting heights. Cost is low; what do you pay for a 69-foot piece of wire and four insulators? On the lower bands the triangle antenna has favorable low-angle directivity. It also operates well with parasitic reflectors and directors and can be used as elements of a phased antenna system.

open triangle

The closed triangle, like the closed quad or delta loop, is a single-band antenna. Operation on other bands is possible using concentric mountings as for three-band cubical quads. The open triangle antenna, when cut for 20 meters provides good bi-directional performance on 10, 15 and 20 meters (see **fig. 14**). Note that the insulator at the apex of the triangle divides the antenna into two half-wave 20-meter sections. The feedpoint impedance is high and requires an antenna tuning unit with both series and parallel tuning.

The open triangle also loads well and gives reasonable performance on 40 meters. On this band each segment, at 34.5 feet, is a bit longer than a quarter wavelength. Although there is a lot of experimentation to be done with the triangle, it is a good base for high-performance, low-budget antennas.

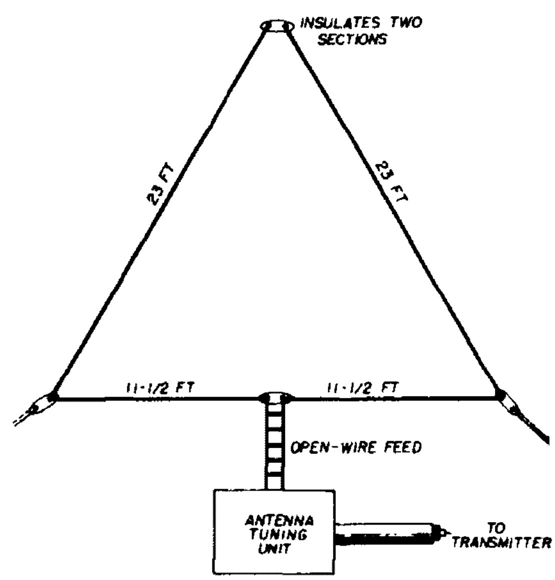
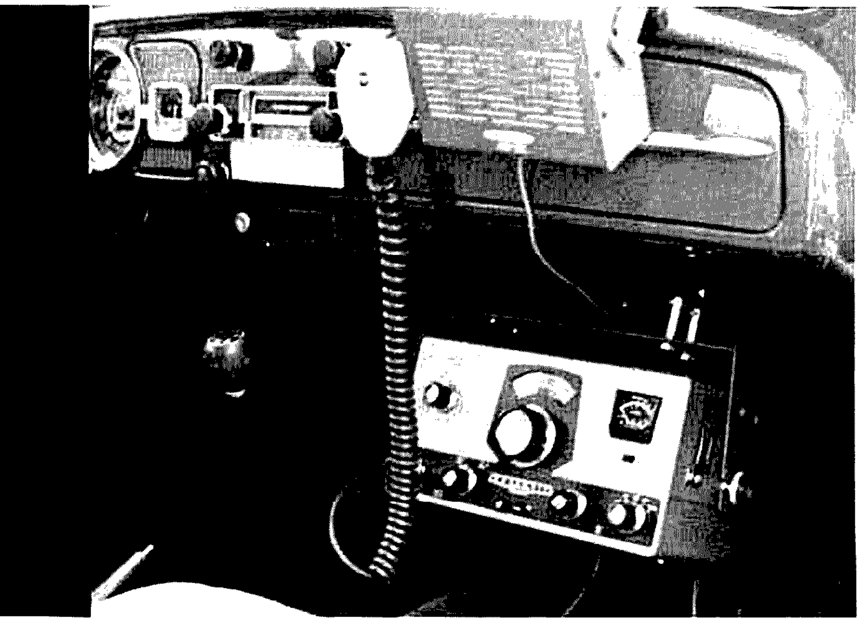


fig. 14. Open triangle antenna cut for 20 meters operates successfully on 10, 15 and 40 meters. The antenna wire is broken at the apex of the triangle.

references

1. Paul Hrivnak, VE3ELP, "Integrated-Circuit Communications Receiver," *ham radio,* July, 1971, page 6.
2. Pat Hawker, G3VA, "An Integrated-Circuit Linear Amplifier," *Radio Communication,* November, 1970, page 761.

ham radio



mini-mobile

It is possible to operate mobile from a small foreign car — this VW mobile is a good example

Donald Backys, K9UQN, 138 Valley Lane, Apartment 205, Hoffman Estates, Illinois

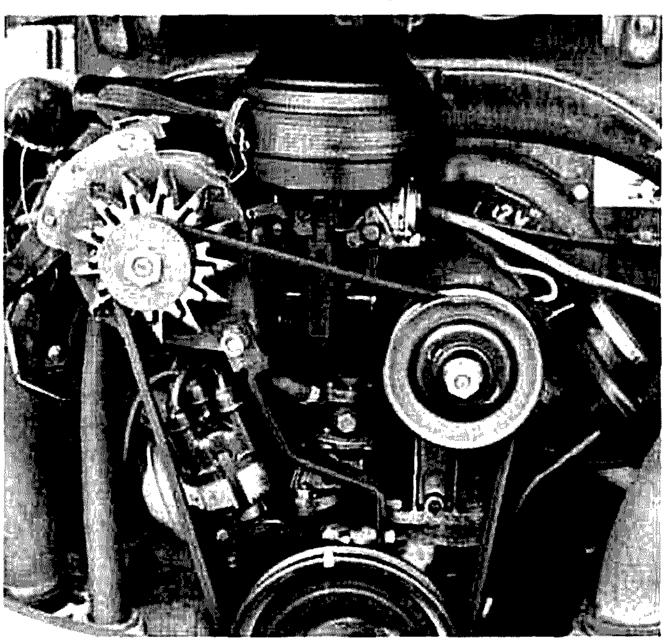
When my wife first suggested buying a Volkswagen for our first car I was somewhat stunned – knowing it would never materialize into the mobile station I always dreamed of. Going against the odds of placing all by mobile gear in the small Beetle I began to change the VW into a rolling radio station. I hope this article will provide incentive to all hams who own small cars to try mobile operation.

The dc power supply is installed up front in the VW trunk.

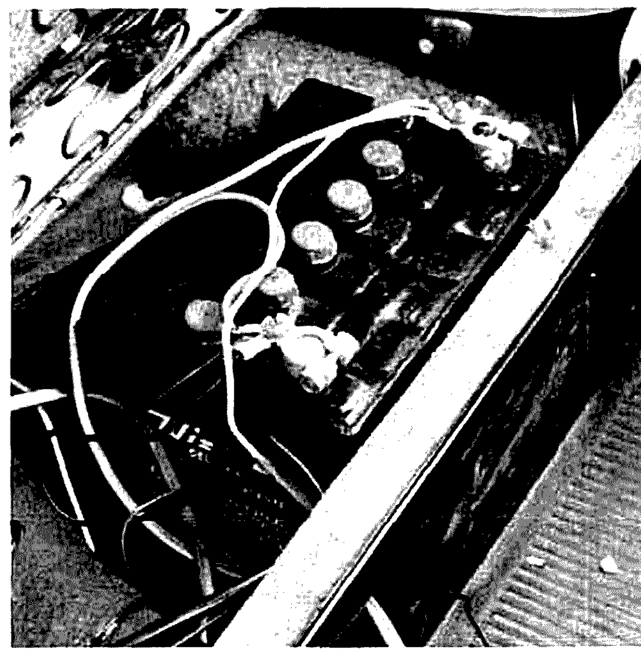
In my mobile radio station a Heath ssb transceiver is mounted under the dash on the passenger's side. It requires some ingenuity in fabricating mounting hardware; however, no special tools are necessary. A standard magnetic mobile microphone holder is mounted on the glove compartment to allow easy access to the microphone with minimum distraction from driving chores. The mobile speaker is mounted next to the microphone. If you wish, the car broadcast radio speaker may be used by placing a toggle switch under the dash that switches the speaker from the car radio to the mobile radio. This is an added advantage in the VW since the speaker is directly in front of the driver.

The dc power supply was mounted in the trunk by cutting an access hole in the trunk liner and fabricating mounting hardware to compensate for the contours of the trunk base. You may question the mounting position of the power supply; however, no problems have been encountered with overheating; even on hot summer days.

A 35-amp alternator takes the load off the VW electrical system.



The heart of the VW mobile, and the key to its success, is its 12-volt power plant. A separate 35-amp alternator system was installed to eliminate the possibility of overtaxing the VW electrical system.* The regulator was installed in the upper left corner of the engine compartment. A separate battery was



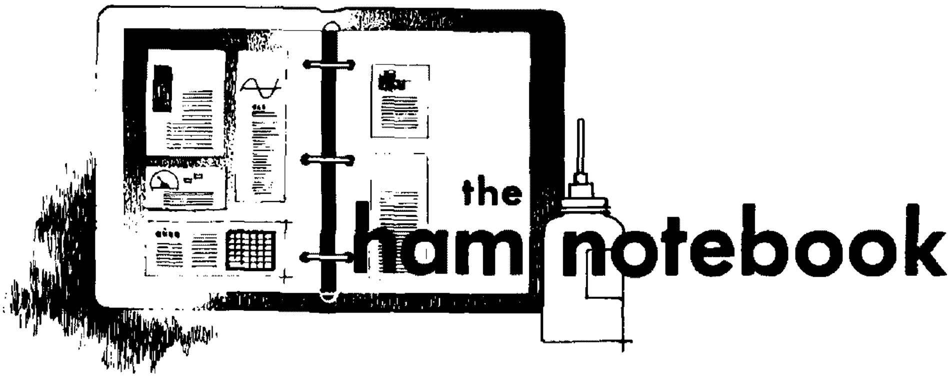
Heavy-duty battery under the rear seat is used exclusively for the radio gear (can also be used for starting the car on cold winter mornings).

installed under the rear seat with battery cables running to the power supply in the trunk.

When I work fellow VW owners on the air they invariably ask me how it was done. It is hoped the ideas presented here will make it possible for other VW owners to have a mobile radio station. For those of you who are wondering about the added drag of the alternator, I found that approximately 3 to 5 mph was knocked off the top speed; however, there was no noticeable decrease in gas mileage.

*Alternator, part number RM 12V PAK, available from 767 Minot Avenue, Auburn, Maine 04210.

ham radio



motorola fm receiver mods

During the past year I have retuned several 43G and 43GGV Motorola Sensicon transceivers for amateur use on two meters. In some of the units I discovered that the squelch control had to be advanced almost fully clockwise to quiet the channel. After checking tubes, components and voltages and a careful alignment I still could not find the problem.

Checking through a number of manuals, I found some resistor changes in the schematic for a late T-Power receiver strip. When these resistor values were incorporated in an otherwise identical older version of the receiver the result was vastly improved squelch-control action. The new resistor values are shown in table 1.

To provide a measure of protection to the 6AQ5 audio output stage I installed a 120-ohm, 1-watt resistor between the cathode (pin 2) and ground, bypassed with a 50-μF, 25-V capacitor. Gain of the stage is not greatly affected, and the tube will be protected by self-bias.

table 1. New resistor values for the Motorola Sensicon G fm receivers.

resistor	old value	new value wide band	new value narrow band
R139	-	56k	180k
R140	-	56k	180k
R145	150k	330k	330k
R146	-	680	220
R152	560k	820k	1.5M

table 2. Squelch and audio-voltage readings are helpful in troubleshooting Motorola Sensicon G fm receivers. (Vrms readings are noise voltage measurements made with ac vtvm.)

test point	unsquelched	squelched
V112, pin 1	3 V rms	3 V rms
V112, pin 2	7 Vdc	2 Vdc
V112, pin 5	50 V rms	50 V rms
	190 Vdc	100 Vdc
V112, pin 6	100 Vdc	60 Vdc
V113, pin 6	12 Vrms	12 Vrms
V114, pin 1	45 Vdc	25 Vdc
V114, pin 2	-8 Vdc	2 Vdc
V114, pin 3	3 Vdc	3 Vdc
V114, pin 6	180 Vdc	190 Vdc
V114, pin 8	50 Vdc	50 Vdc

Table 2 lists a number of dc voltage readings at various points in the squelch and audio stages. These should be helpful in tracking down circuit ailments.

Murray Ronald, VE4RE

multiple tubes in parallel grounded grid

The use of four parallel tubes in grounded-grid dates back many years. If the tubes are honest-to-goodness triodes like the 811A they work pretty well. However, if you use tv-type horizontal-amplifier tetrodes or pentodes, tie all of the grids together and ground them, the picture is a little different. When you get that much grounded hardware in close proximity to the plate of the tubes the plate-to-ground capacitance becomes formidable, particularly when you have four tubes in parallel.

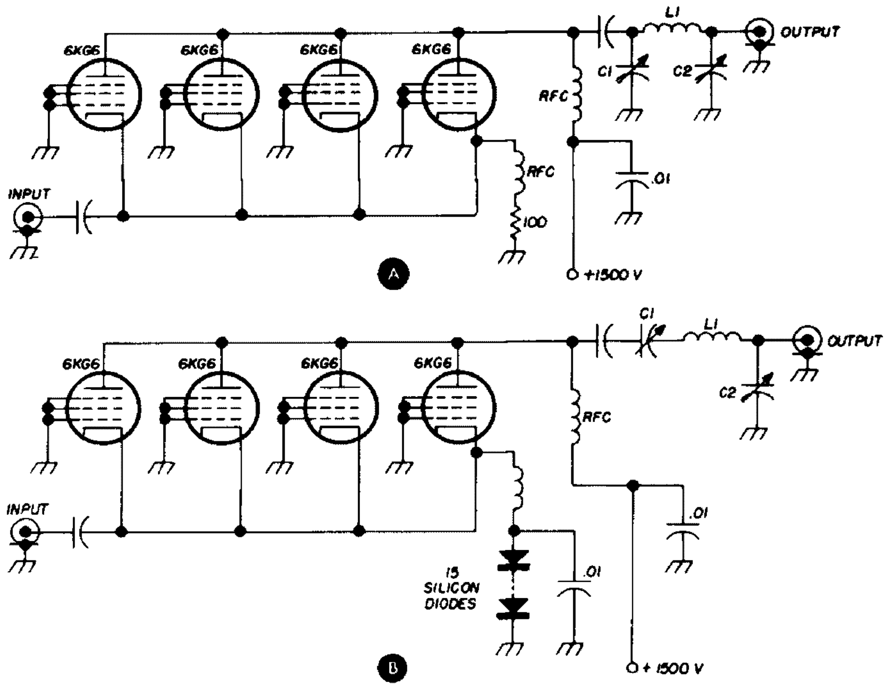


fig. 1. High capacitance of parallel-connected tv sweep tubes limits upper frequency capability of the amplifier circuit in (A). Series capacitor C1 in (B) improves L/C ratio of the circuit and provides increased high-frequency efficiency.

Several years ago I built such an amplifier using four 6KG6 Amperex tubes. These tubes are somewhat huskier than the 6LQ6 and harder to come by, but they will take a lot of abuse. I have run the four of them to a full 1000 watts dc. One tube finally blew so power was dropped to a conservative 500 watts dc (1000 watts PEP).

At the lower frequencies the standard grounded-grid circuit shown in **fig. 1A** worked fine. However, at the higher frequencies the plate-to-ground capacitance is on the same order of the tuning capacitor and in parallel with it. Result: poor efficiency on 15 meters and impractical on 10 meters.

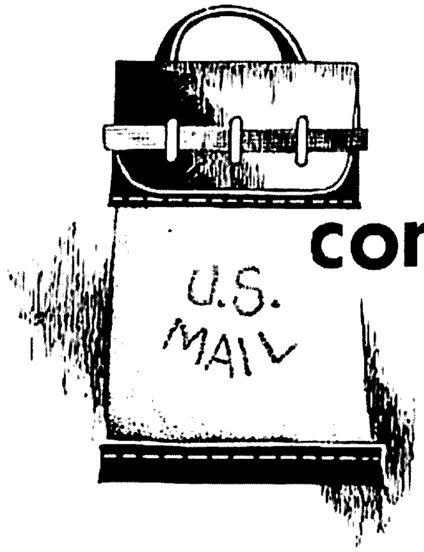
Recently I decided that I wanted to use this amplifier on 10 and 15 meters, but first I had to solve the high-C dilemna. The solution is shown in **fig. 1B.** In **1B** the input pi-network capacitor C1 is placed in series with the plate-to-ground capacitance. With this arrangement capacitor C1 becomes the series combination of the plate-to-ground capacitance of the four tubes *and* the variable tuning capacitor. You have to insulate both sides of the tuning capacitor from ground now but this is no big problem.

With circuit **1A** on 15 meters the tank coil consisted of two very hot turns; with fig. **1B** the tank circuit has 6 turns and everything runs cool, as it should. I drive the amplifier with an SB-34 transceiver.

One other major improvement was incorporated into the rebuilt amplifier. The original circuit used a 100-ohm cathode-bias resistor (**fig. 1A**). This is a poor way to bias a variable-plate-current power amplifier. The bias circuit in **fig. 1B** uses fifteen series-connected silicon rectifiers to obtain a 10-volt drop which is *constant* and limits the resting current to about 60 mA. The cost of this arrangement was kept down by using surplus 100 PIV 1500 mA silicon diodes that are available from Poly-Paks for less than a dime apiece. A more expensive approach to the bias problem is a 10-volt, 20-watt zener diode.

The rebuilt linear delivers a very clean signal. Although I used 6KG6 tubes the same technique may be used to solve the high plate-to-ground capacitance of any parallel-connected tv sweep tubes.

Bob Baird, W7CSD

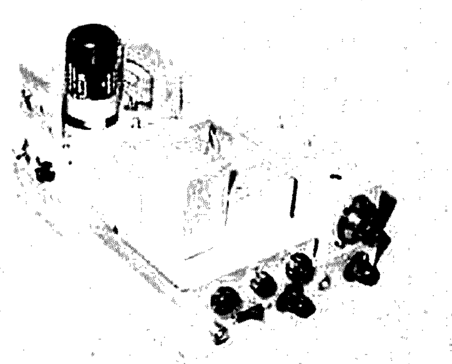


comments

solid-state ssb exciter

Dear HR:

I just wanted to let you know that some of us actually build equipment described in *ham radio.* I'm enclosing a photo of a ssb exciter I came up with after reading W9KIT's article in the December issue. The basic 9-MHz section is basically the same as that described in the article, although I haven't put the crystal filter in yet.



In correspondence with the author I got the dope on his entire ssb exciter, and followed his heterodyning scheme pretty much, although I used transistors and he used vacuum tubes. The double-balanced mixers work out nicely; although this is my first experience with them I remember an article you ran on these gadgets some time ago.

It sure is a kick to work somebody on phone and tell them I'm using a homebrew rig. Most of the guys are never aware that I'm using double sideband unless I tell them, so I may never get around to putting in the crystal filter as I'm primarily interested in cw anyway! This little rig drives my 6146 final to about 60 watts PEP input which seems to be plenty to make contacts on 40-meter ssb.

Cal Sondgeroth, W9ZTK
Mendota, Illinois

fire protection

Dear HR:

In the article "Fire Protection in the Ham Shack" in the January 1971 issue, Freon is listed as being a good fire extinguisher. Possibly it would be good for putting out the fire, but might hospitalize the fireman. When subjected to temperatures corresponding to that of red-hot metal, Freon breaks down, forming Phosgene gas, which my dictionary describes as a colorless, highly toxic gas; one use of which is chemical warfare. As personal experience followed by research has taught me, phosgene is a lung irritant, also an eye irritant and very good for rusting tools. The authors are not to be blamed too harshly. Freon is non-toxic, stable and non-inflammable — but only at ordinary temperatures.

Frank Knottingham, K7QCM
Gold Beach, Oregon

A set of schematics for W9ZTK's solid-state ssb exciter for 80, 40 and 20 meters is available for 25c from *ham radio.* The circuit includes a 9-MHz sideband generator, speech amplifier, vfo, double-balanced mixers for 80, 40 and 20 meters plus an fet output amplifier. *editor*

new products

frequency standard

A zero adjustment allows the frequency to be set precisely to WWV or other frequency standard. The accuracy holds for all outputs of the frequency standard. Applications include calibration of receivers, signal generators, vfos, oscilloscopes and general use in the lab or ham shack. The unit is priced at $32.50 postpaid from Palomar Engineers, Box 455, Escondido, California 92025. For more information use *check-off* on page 94.

The new Palomar Engineers frequency standard contains a crystal-controlled oscillator that is based on a precision 100-kHz crystal. A front-panel switch controls the operation of three frequency dividers driven by the 100-kHz oscillator. They provide outputs at 100, 50, 25, 10 and 5 kHz, depending on the switch setting. The output is a square wave with rich harmonic content; thus the signal can be heard at every 100 kHz (or every 50 kHz, or every 25 kHz, etc.) throughout the shortwave bands to 50 MHz and above.

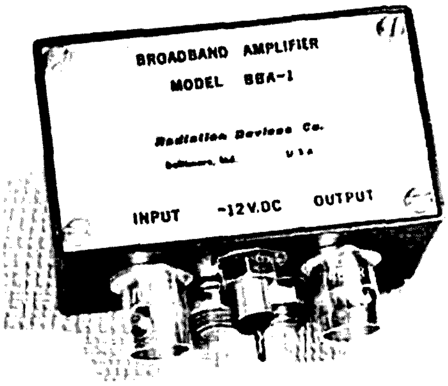
afsk generator

The new J&J solid-state afsk generator provides short-term stability of ±1 Hz. The basic oscillator uses a unijunction transistor in an RC circuit with precision resistors. The internal 5-pole butterworth filter with cutoff above 3000 Hz removes harmonics and assures sinusoidal output.

The J&J afsk generator features high output with more than ample transmitter drive. The high level line output may be used to test TUs or other RTTY equipment. The generator may be triggered by any positive voltage greater than 2 volts. A keyboard can be connected to the input of the generator with a power supply up to 20 volts to cause the unit to conduct on *space*. The afsk generator can also be operated directly from any of the TT/L or equivalent keying circuits which supply a positive voltage on *space*.

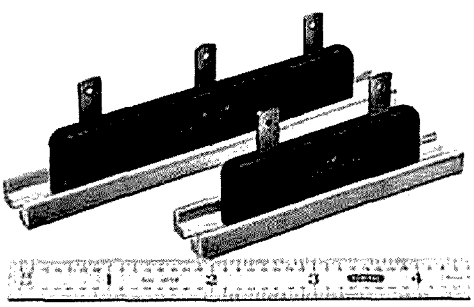
The J&J afsk generator is furnished with standard tones 2125, 2295 and 2975 Hz unless otherwise specified. Price, in modern desk-top cabinet, is $90.00; $99.00 for rack-mounted model (3 1/2" panel). Also available with low tones of 1275, 1445 and 2125 Hz at $99.00. Order from J&J Electronics, Windham Road, Canterbury, Connecticut 06331. For more information use *check-off* on page 94.

broadband rf amplifiers



The new broadband rf amplifiers introduced by Radiation Devices Company span the region from 10 to 500 MHz for state-of-the-art performance for laboratory or field use. Gain is 30-dB *minimum* from 10 to 150 MHz; gain from 3 to 500 MHz is 10 dB for the model BBA-1 and 15 dB for the BBA-1P. Noise figure, from 10 to 500 MHz, is 6 dB maximum for the BBA-1, and 3 dB maximum for the BBA-1P. The broadband amplifiers require a supply voltage of 12 volts, ±1 volt. Input and output connectors are type BNC (other connectors available). The BBA-1 is priced at $30; the BBA-1P is priced at $45. For more information, use *check-off* on page 94, or write to Radiation Devices Company, Post Office Box 8450, Baltimore, Maryland 21234.

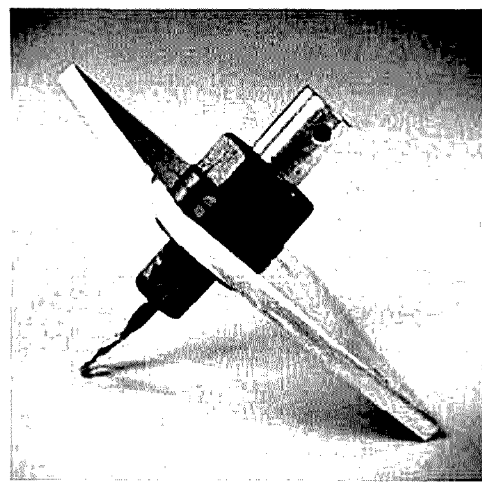
high-voltage silicon rectifiers



Semtech has announced the development of a unique high-voltage silicon rectifier device, called the KV-PAC, with peak inverse voltages up to 15 kV. The KV-PAC offers a rugged functional pack-

age that is corona free. Mounting slots are designed for easy installation. Universal 3-watt electrical terminals accommodate fast-disconnect fittings, wire wrap or solder connections. The devices are rated at 5 to 15 kV PIV per leg. Average rectified current is 0.4 amp; one-cycle surge current is 50 amps. Reverse current at the rated PIV is 10 μA. For more information, use *check-off* on page 94, or write to Semtech Corporation, Newbury Park, California 91320.

stud-nut

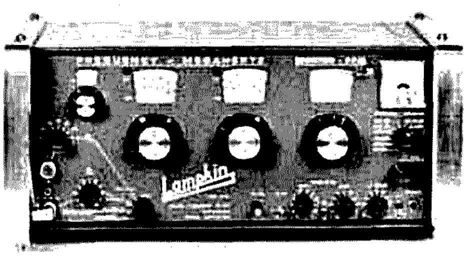


The *Stud-Nut* is a new unitized fastener which eliminates five separate pieces of hardware now required to mount stud-type solid-state devices. It replaces an insulating bushing, mica washer, metal washer, solder lug and hex nut; it provides rapid, secure mounting of SCRs, power transistors, rectifiers and zener diodes with 10-32 threaded studs. It can be used with DO-4, TO-36, TO-59, TO-60 and TO-64 case types.

The threaded metal insert is molded into an insulated hex base and contains through holes to provide solder connections for external leads. Due to the unique design of the *Stud-Nut* the contact resistance between the stud-mounted device and external lead is extremely low. Continuous current rating is 30 amperes. For more information, use *check-off* on page 94. A package of four *Stud-Nuts* is available for $1.00 postpaid from the SCF Corporation, Post Office Box 999, Hightstown, New Jersey 08520.

More Details? CHECK—OFF Page 94

digital frequency meter/ signal generator



The recently announced type 107A digital frequency meter/signal generator/ synthesizer from Lampkin Laboratories is an extremely versatile instrument that is aimed primarily at the mobile-radio maintenance field. The instrument has many applications in educational, aerospace and industrial laboratories; in manufacturing, research, and production operations; in a-m, fm, and tv broadcast engineering and receiver servicing; in commercial frequency-measurement services and in many other applications. It is completely solid-state, operates from either 12 Vdc or 115 Vac and weighs 22 pounds.

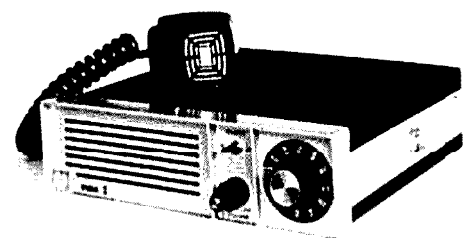
As a heterodyne frequency meter, the 107A will measure carrier frequencies of nearby transmitters or signals picked up on a receiver. Coverage on FCC-assigned frequencies is continuous from 10 kHz to above 500 MHz. Guaranteed accuracy in the field, independent of WWV, is considerably better than 0.0001% (1 part per million).

As a synthesizer any frequency from below 1,000 Hz to 9,999.9 Hz, can be generated, in steps of 100 Hz, phase-locked to the internal crystal standard. Voltage output level is 1.0-volt rms down to 0.0005-volt, continuous and calibrated, into 50-ohm load or greater. As a signal generator the 107A provides cw, amplitude or frequency-modulated signals on fundamental frequencies up to 10 MHz; from 10 MHz to above 500 MHz harmonics are employed, but no arithmetic is needed – the dials are direct reading in MHz and kHz. An external variable attenuator (supplied) will bring the signal below the noise level of receivers.

More Details? CHECK–OFF Page 94

Price of the Lampkin 107A is $2150.00. For further information, write Lampkin Laboratories, Inc., Dept. 243, 8400 Ninth Avenue N. W., Bradenton, Florida 33505.

vhf marine radio



Every marina or yacht club, or any marine service company, can now afford its own VHF Limited Coast Station with the Ensign II produced by RF Communications. Ideal for marine sales and service dealers to have as a unit to show, demonstrate and use, all at the same time, the Ensign II provides full 25-watt output. Transceiver weight is under 7 pounds, and the unit includes crystals aligned for channel 16 and a working channel ready for use. Capable of being mounted in any position, the Ensign II is housed in a vinyl painted cabinet with drip-proof and dust-tight construction. List price is $578.

A complete line of antennas and accessories is also available. For literature with full specifications and further information use *check-off* on page 94, or contact National Sales, RF Communications, Inc., a subsidiary of Harris-Intertype Corp., 1680 University Avenue, Rochester, New York 14610.

wwv changes

When you tune in WWV (or WWVH) for time signals or propagation information after 0000 hours gmt, 1 July 1971, you will hear a different format than you've been accustomed to over the last few years. The Morse code transmissions will be gone and the announcements of time and other information will be made in voice. The time will be announced

More Details? CHECK—OFF Page 94

every minute instead of every five minutes; a male voice will be used by WWV and a female voice by WWVH to distinguish between them. The carrier frequencies will remain the same: 2.5, 5.0, 10.0, 15.0, 20.0 and 25.0 MHz. The frequency accuracy of the two stations is controlled by the NBS Atomic Frequency Standard at Boulder, Colorado.

The format of the broadcasts from the two stations will be similar, but to avoid confusion they will use alternate time slots for the transmission of tones, announcements, etc. The standard time tick each second will remain.

Each hour will be divided into 1-minute slots. Each minute (except the first) will begin with an 0.8-second 1000-Hz tone on WWV (1200 Hz at WWVH). The first minute of each hour will begin with a 1500-Hz tone at both stations. The one minute slots will be divided into a 45-second segment and two 7.5-second segments. On alternate minutes the 45-second segment will contain either a standard tone or an announcement. The first 7.5-second segment following the 45-second segment will be used by WWVH to announce time while WWV will be silent. The second 7.5-second segment will be used by WWV to announce time, WWVH being silent.

Each station will omit for 5 minutes of each hour all tones and announcements during the 45 second segments. This period will begin for WWV at 45 minutes past the hour and for WWVH at 15 minutes past the hour. A special 440-Hz tone will be broadcast by WWV for 45 seconds beginning one minute past the hour and by WWVH two minutes past the hour. This tone can be used to mark the hours on strip-chart recorders or other instruments. The tone will be omitted during the zero hour (gmt) of each day.

It is interesting to note that announcement slots will be available to Government agencies to use for their own purposes. The slots not used will be filled by a standard tone, probably 500 Hz. To prevent interference between the two stations one station's announcements will coincide with the other's tone.

75 cents

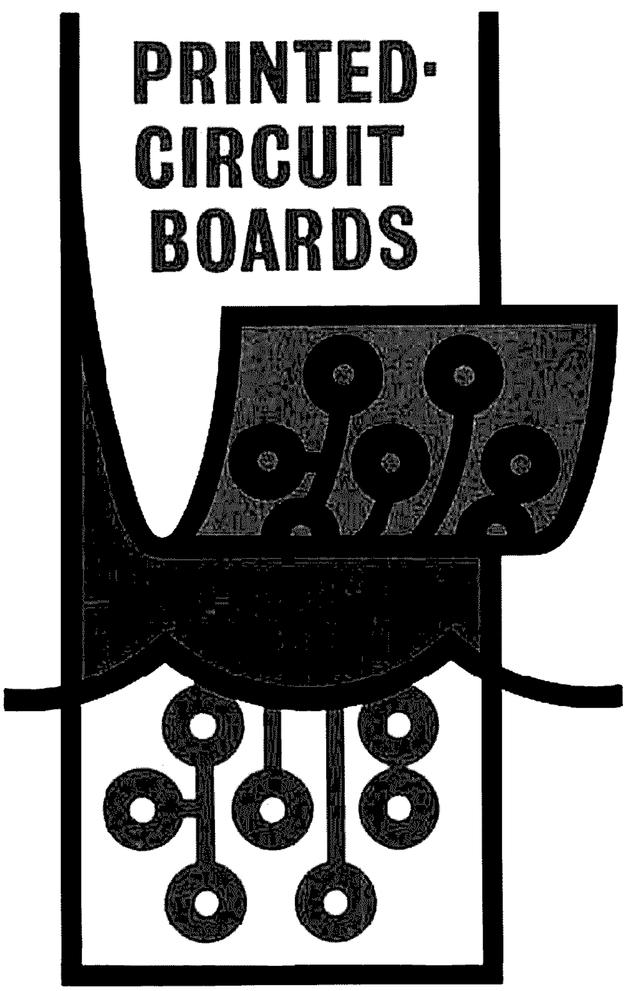
hr

focus on communications technology . . .

ham radio magazine

SEPTEMBER 1971

PRACTICAL PHOTOFABRICATION OF



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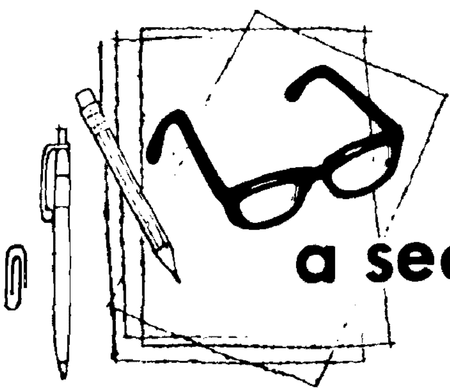


a second look

by jim fisk

A new electronically-tunable ceramic capacitor has been developed by engineers at Fort Monmouth, New Jersey. The new tunable capacitor is about one-tenth the size of presently available units, equivalent to three quarters stacked on top of one another.

The tunable capacitor, which will be used in automatic tuners in lightweight military radios, costs about thirty cents. The device consists of a multilayered ferro-electric material with an rf electrode on each end and a dc electrode in the center. As the dc control voltage is varied, the dielectric constant of the center layer changes, thereby changing the capacitance of the device.

The voltage-tunable characteristic of these new ceramic capacitors is particularly valuable for applications which require automatic tuning capability. For example, the capacitor can be combined with a digital logic network to retune an antenna automatically if a mismatch occurs between it and the transmitter. In a typical application the radio can completely retune itself automatically within several microseconds.

Varactors are not suitable for this application because of their power limitations. To obtain high power, a larger number of series-connected diodes must be used; this lowers capacitance. To increase capacitance, more diodes must be placed in parallel; this results in a very unwieldy package. With the ceramic multilayer capacitor, however, a whole stack of devices can be placed in a very small package.

Since these multilayer capacitors are still in the developmental stage, it will be some time before they become available for use in radio production. Eventually they will reach the public. When they do we should see a whole new array of automatically tuned communications antennas.

For example, by including one of these tunable ceramic capacitors inside the traps of a multiband beam, an enterprising amateur could build a high-performance antenna system which would provide unity swr at any operating frequency. The required digital logic control circuitry could be built into the antenna boom.

These ceramic capacitors could also be used in automatic tuning units at the output of your transceiver, or as tuning and loading controls within the equipment. The capacitors would also be ideal for automatically tuned mobile antennas which compensate for swr changes as you drive under trees or next to large tractor-trailors. This is especially important when using solid-state final amplifier stages which quickly destroy themselves when confronted with high standing-wave ratios.

Jim Fisk, W1DTY
editor

practical photofabrication of printed-circuit boards

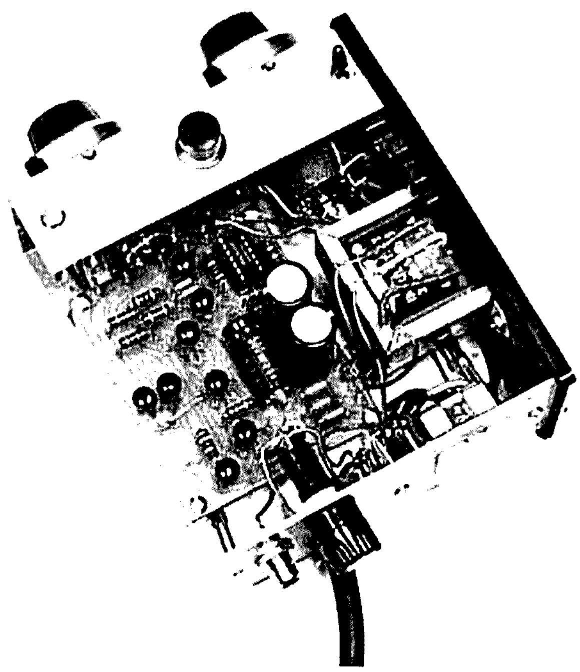
Here's a practical inexpensive technique for producing professional quality printed-circuit boards for your amateur radio projects

Larry Hutchinson, Chemistry Department, Oregon State University, Corvallis, Oregon 97331

Making the printed-circuit board for a project is often as much fun as using the finished project itself. The photoresist technique is easy, inexpensive and results in a neat, professional looking and aesthetically pleasing final project. A quick scan of the literature reveals that this technique has actually been used very rarely by amateurs. This is because of the apparent unavailability of photoresist, misconceptions about cost and, of course, the absence of a comprehensive article on the subject.

The method of making printed-circuit boards detailed in this article consists of making a transparency using readily available drafting aids and coating a piece of

Completed project using the printed-circuit board.



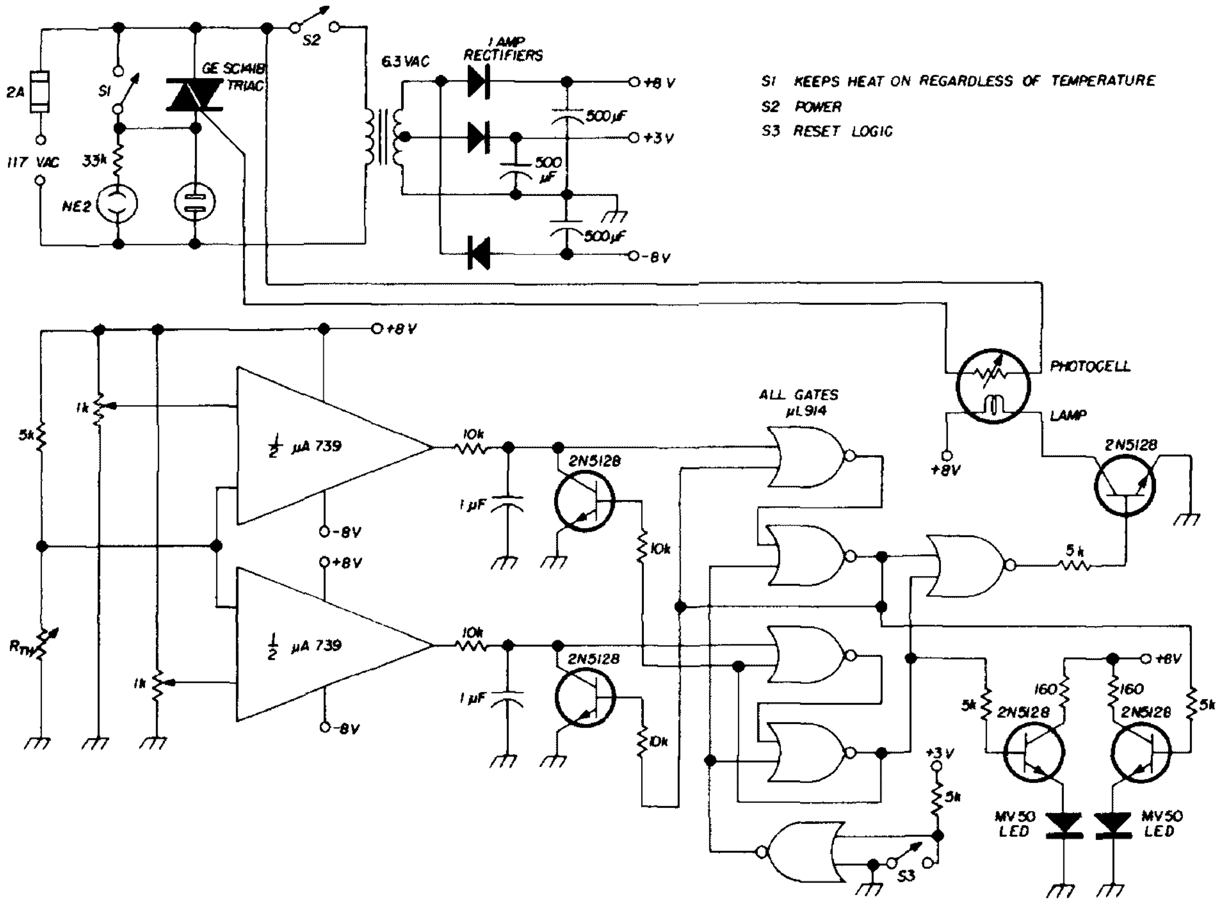


fig. 1. Schematic of the automatic-shut-down circuit used to illustrate the photoresist method of making printed-circuit boards.*

copper-clad board with a positive-working photoresist, exposing, developing and etching. A conservative estimate of the photoresist cost for a 5 x 7-inch board is $.10. No darkroom, camera or photographic skill is required since the only light-sensitive part of the process is the photoresist, and it is sensitive only to ultraviolet light. A photoresist-coated circuit board may be exposed to normal room lighting for an hour or longer with no harm.

*This equipment is used in an organic chemistry laboratory to automatically stop a distillation when the temperature of the material being distilled either rises above the boiling point of the pure material (indicating the presence of a higher boiling impurity) or falls below the boiling point (when most of the material is gone from the pot and not enough hot vapor is flowing past the thermistor in the distillation head to keep it hot). The Triac in the circuit is used as an on-off switch to determine whether heat is applied to the pot (through a heat controller plugged into the socket). The Triac is controlled by the logic through a lamp/photocell combination. The light-emitting diodes are used to indicate whether the distillation was shut-down due to a rise or a fall in temperature.

Techniques for making printed-circuit boards may be divided into two basic categories. Those methods characterized by application of resist to areas where copper remains after etching, and methods using photoresist or silkscreen.

The first method is useful for one-of-a-kind projects. Examples include paint, ink, tape and dry-transfer images applied directly to the circuit board. Also included is stick-on copper foil with heat-resistant adhesive; this is a very interesting material since etching and drilling are eliminated, and the circuit can be changed with relative ease.

Photoresist methods are divided into

three types: those that use a positive-working photoresist with a positive transparency, develop and etch; those that use a negative-working photoresist with a tive photoresist is Shipley's AZ-111 which is developed in a weak solution of sodium hydroxide (lye).

Negative-working photoresists consist

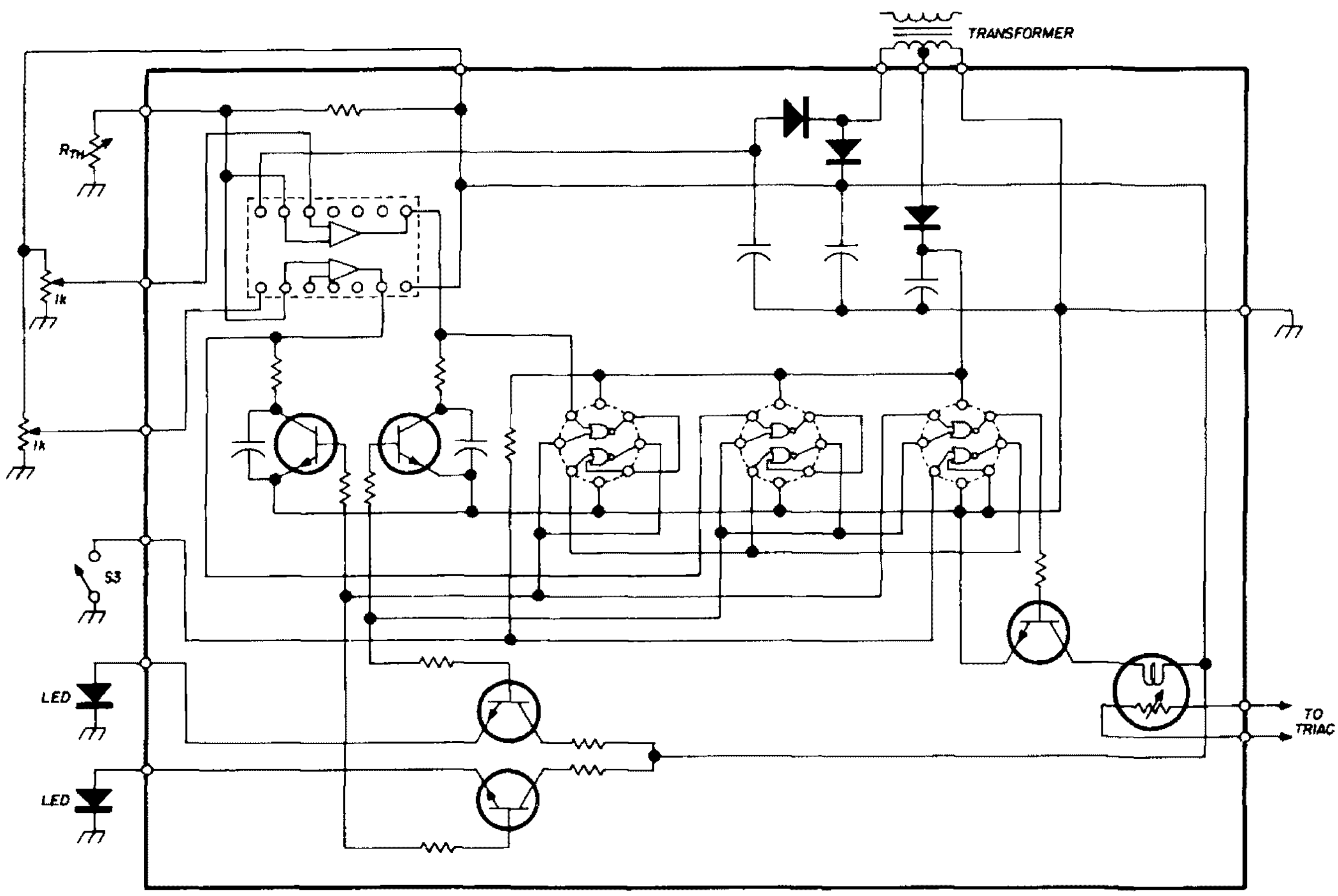


fig. 2. First step toward a printed-circuit board consists of redrawing the circuit, placing components within the proposed board outline. No attempt is made to minimize wiring crossovers at this point.

positive transparency, develop, tin plate, strip resist and etch with an etchant which will not dissolve tin; and those that use a negative-working photoresist with a negative mask, develop and etch. Although only the first method is recommended all three will be covered.

photoresists

A positive-working photoresist consists of an organic resin which is affected by ultraviolet light in such a way that the exposed resin is easily dissolved in its developer, but the unexposed material is not. Positive-working resists are resistant to pinholing and other effects caused by dust particles and other contaminates on *the transparency. An example of a posi-*

of an organic resin which polymerizes when exposed to ultraviolet light such that the resin becomes insoluble in a solvent in which the unexposed resin is easily dissolved. An example of a negative-working photoresist is Eastman Kodak KPR-2; this is developed in trichloroethylene, a common dry-cleaning solvent.

technique

To illustrate the techniques involved in producing a printed-circuit board from scratch I have photographed the various stages in the evolution of a recent project. **Fig. 1** shows the schematic of the circuit; note that the ground and power supply leads are *not fully drawn in and that* the

gates and op amps are represented by symbols which have minimal resemblance to the actual ICs.

The first step is to redraw the schematic according to the following rules:

1. Draw a rectangle to represent the pc board. Parts which will be external to the board should be drawn outside the rectangle in the correct relative positions.

2. Completely draw in power supply and ground leads. Do not bother to minimize lead crossing at this point.

3. Redraw the integrated circuits as shown in fig. 2.

For Step 2 you need to know the size and shape of the printed-circuit board. To do this I first install and wire all the components not on the printed circuit board (see fig. 3). Then it is an easy matter to determine the maximum size and shape of the board, shown as the rectangle in fig. 4.

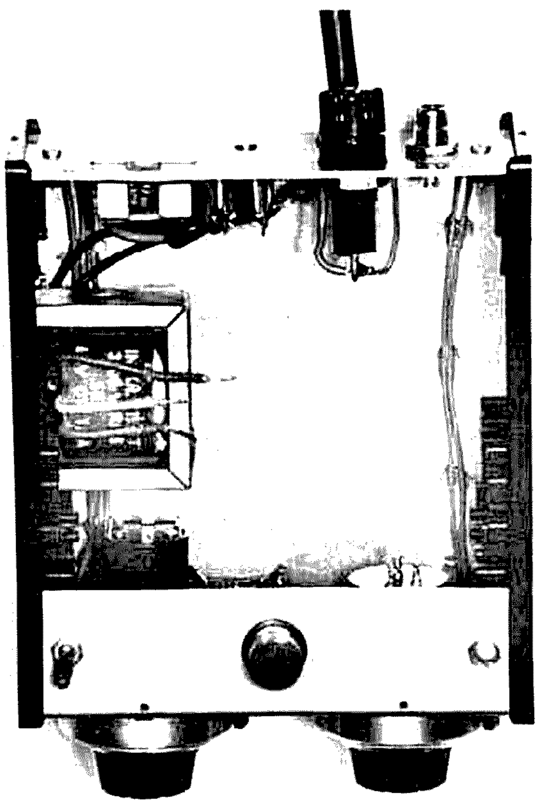


fig. 3. The size of the printed-circuit board and the number of components that it will carry are determined by selecting the enclosure for the project and wiring in those components that will not be on the circuit board.

If you have the option, choose a board which is just large enough to allow horizontal mounting of resistors and capacitors. If you have not done enough bread-boarding of your project and are uncertain as to the workability of the design leave a little extra room for circuit modifications and additions.

The locations of major components such as capacitors, IC sockets and transformers are now determined. Keep in mind the positions of the external components. Pencil in an outline of the large items.

Step 3 involves redrawing the schematic, replacing all symbols with full-size outlines of the components (see figs. 5 and 6). This must be done accurately; the use of graph paper with 0.1-inch grids is helpful in maintaining accuracy. The actual components should be on hand during this step so you can determine if all the parts will fit correctly.

Use pencil. Unless you are working on a particularly simple circuit board you will probably do a lot of erasing. A set of colored pencils is a big help here. The finished drawing (see fig. 6) should be accurate with regard to component positions and all drill holes. When you are certain everything is correctly placed go over the appropriate pencil marks with a

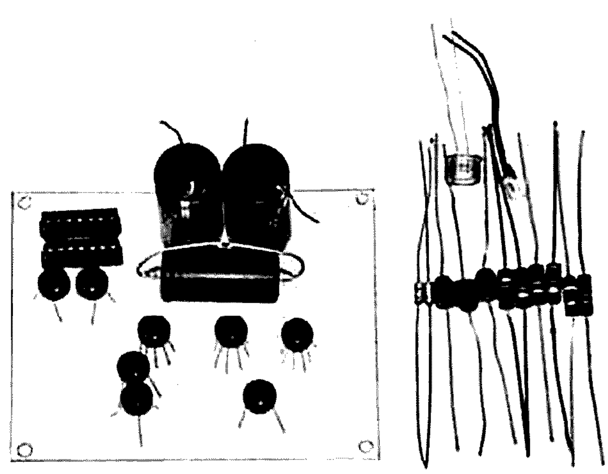


fig. 4. Circuit-board layout is further defined by placing the actual components on a graph-paper outline of the proposed board.

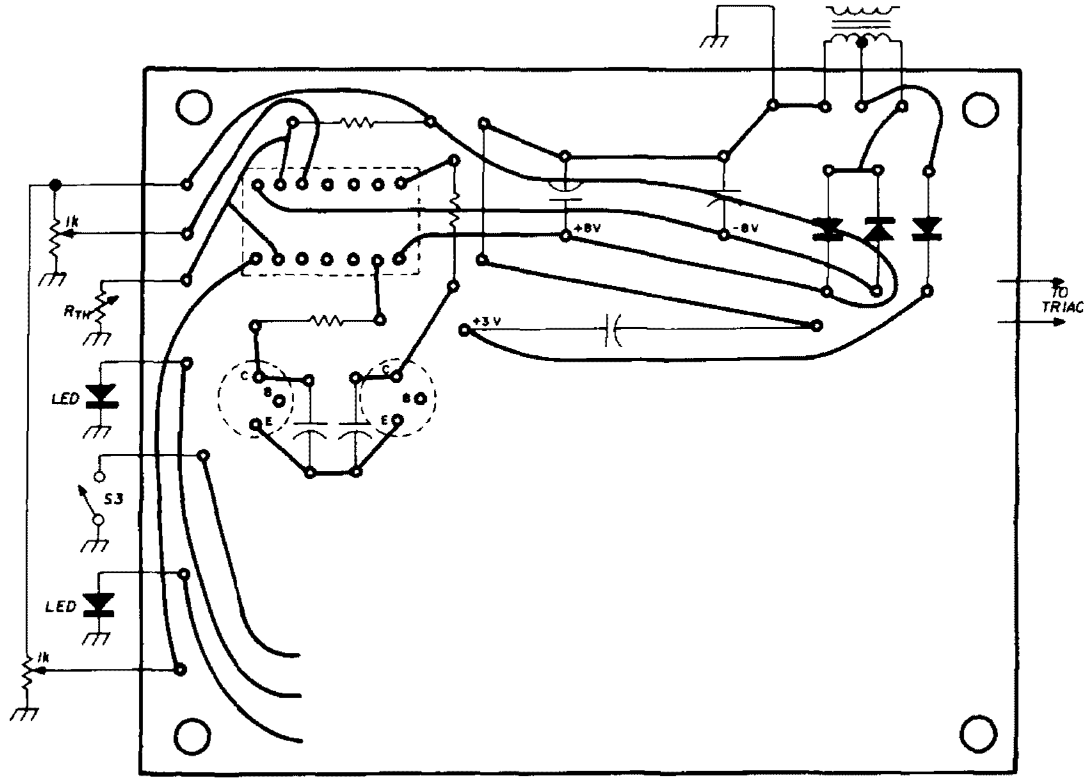


fig. 5. After choosing an initial layout begin sketching in the interconnecting wires. Avoid wiring crossovers.

red felt tip marker so the proposed foil pattern will show through in the next step (see **fig. 7**).

Since **fig. 7** was drawn as a top view the paper must be turned over before making the transparancy. A thin sheet of acetate or mylar plastic is cut a little larger than the board and attached to the back of the diagram with masking tape (**fig. 8**).

the transparency

At this point you will need some of the electronic drafting aids listed in **table**

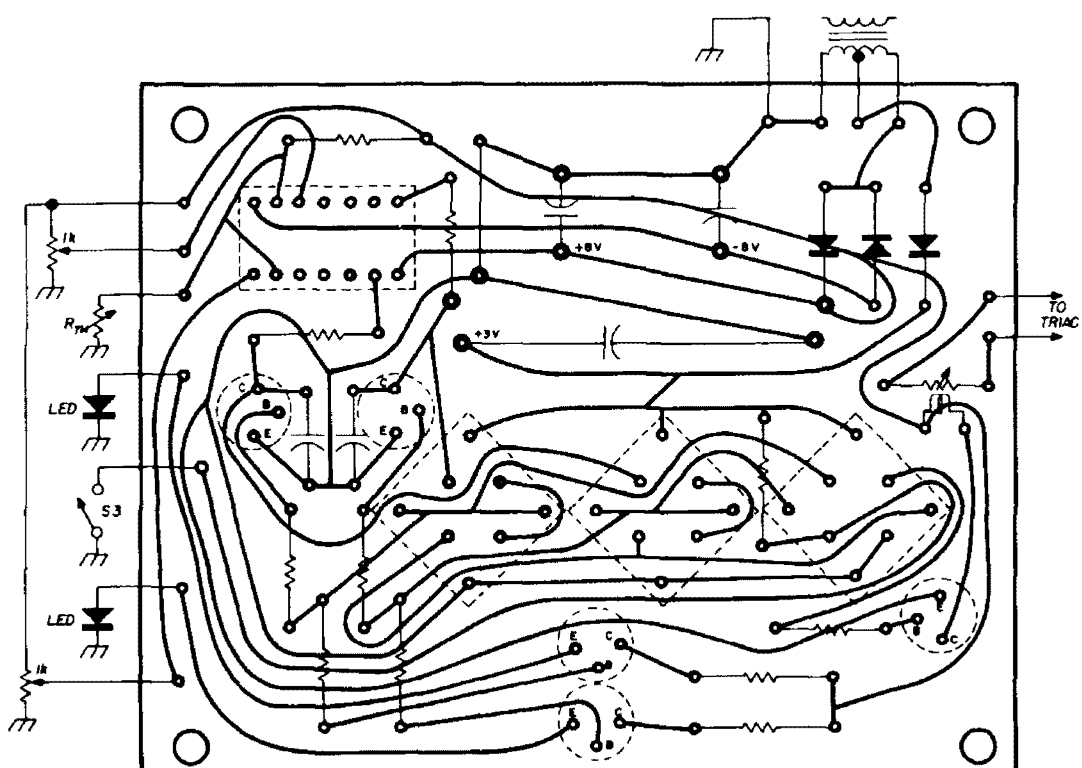


fig. 6. Continue sketching the rest of the circuit layout. Use a pencil — you will probably have to do a lot of component re-arranging to eliminate crossovers.

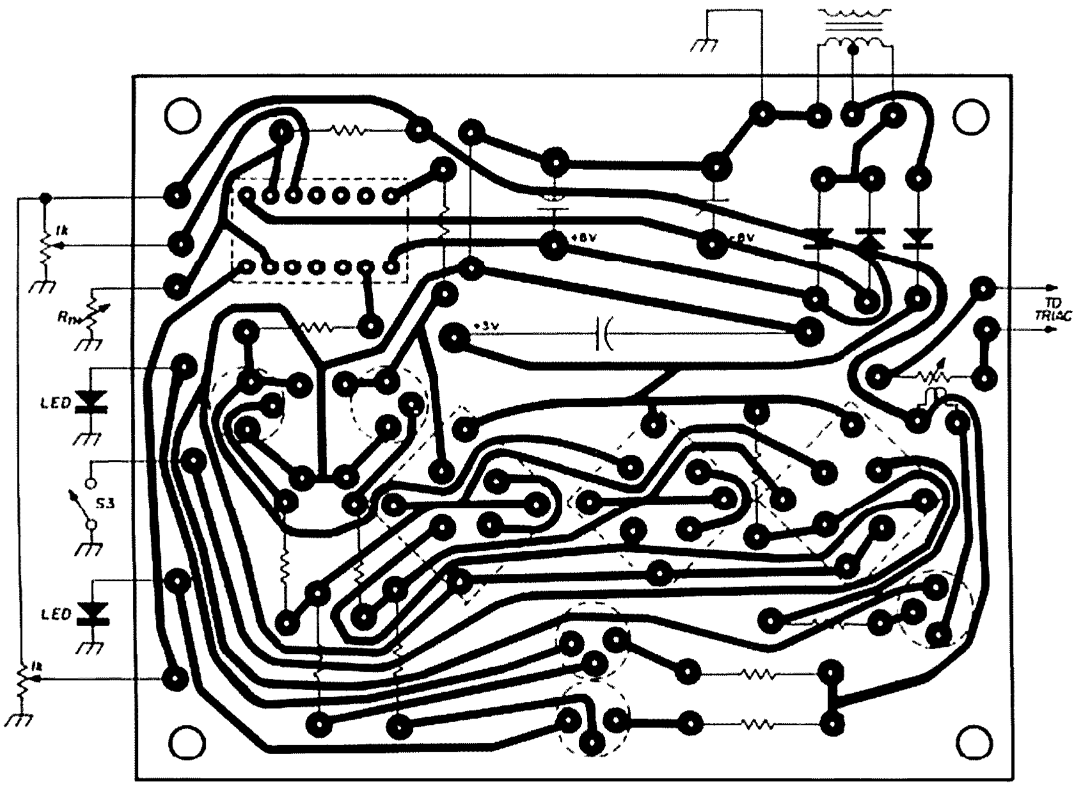


fig. 7. When you are satisfied with the layout go over each of the circuit traces with a dark marking pen.

1. A few of these drafting aids are shown in the photograph. Bishop drafting aids are available in large quantities from Bishop Graphics, and in small quantities from GC Electronics. It is most economical to purchase the most often used drafting aids (tape and donut-shaped pads) from Bishop Graphics, Trans-pak (Chartpak) or Datak, and the less often used aids from GC Electronics. The dry-transfer drafting aids available from Datak are easier to use than conventional die-cut patterns and should not be overlooked.

The drafting aids are transferred to the plastic sheet with a small X-acto knife.

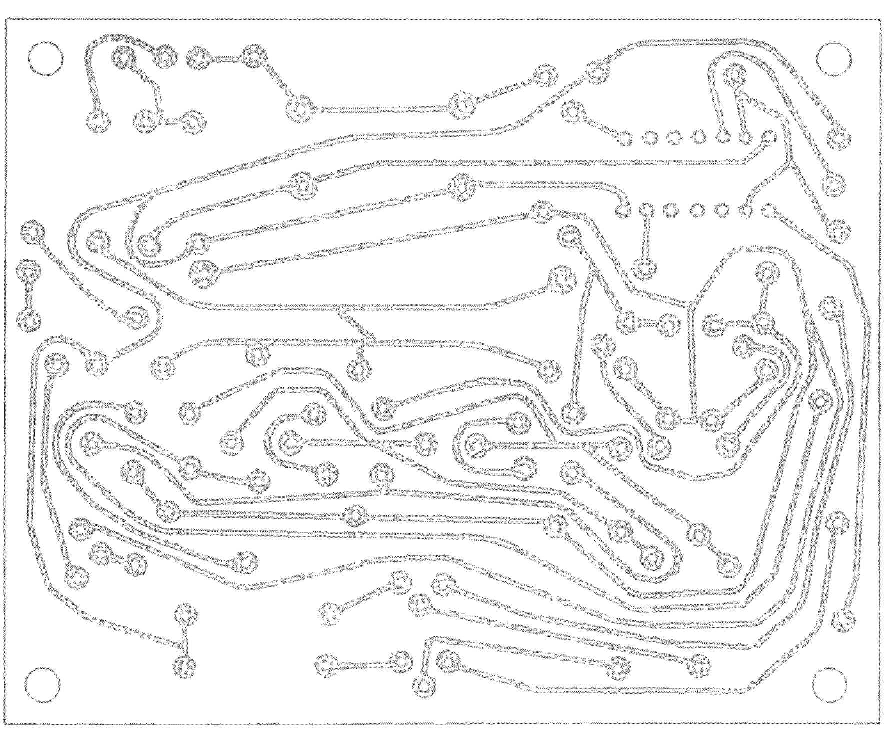


fig. 8. Flop the drawing over; the marking pen outline of the traces will show through the paper.

The donut-shaped pads and similar images should be applied first (fig. 9) followed by the tape (fig. 10). The tape is cut by pressing gently down on the knife, then pulling the free end of the tape up and to the side. It takes a little practice to cut only the tape and not what is below. When you finish the transparency inspect it carefully to see that it corresponds correctly to the circuit. If you are using a negative photoresist a contact exposure with sheet film will produce a negative transparency.

photoresist application

Cut the board to size and sand down any rough edges. The surface of the board should be scrubbed with scouring powder until all oil and fingerprints have been removed. The surface is then washed with a 5-10% solution of sulfuric or hydrochloric acid, rinsed thoroughly with running water, and dried with a paper towel.

The clean board should be coated with a thin, even layer of photoresist within an hour of cleaning. There are many ways this is done commercially including roller coating, spray coating and controlled withdrawal. However, for the amateur, spin application is the best.

Spin application consists simply of applying a generous amount of photoresist to the board and spreading it over the copper surface with a medicine dropper (fig. 11) then spinning the board at 500 to 2000 rpm for a few seconds.

Drafting aids used in preparing the artwork for the printed-circuit board.

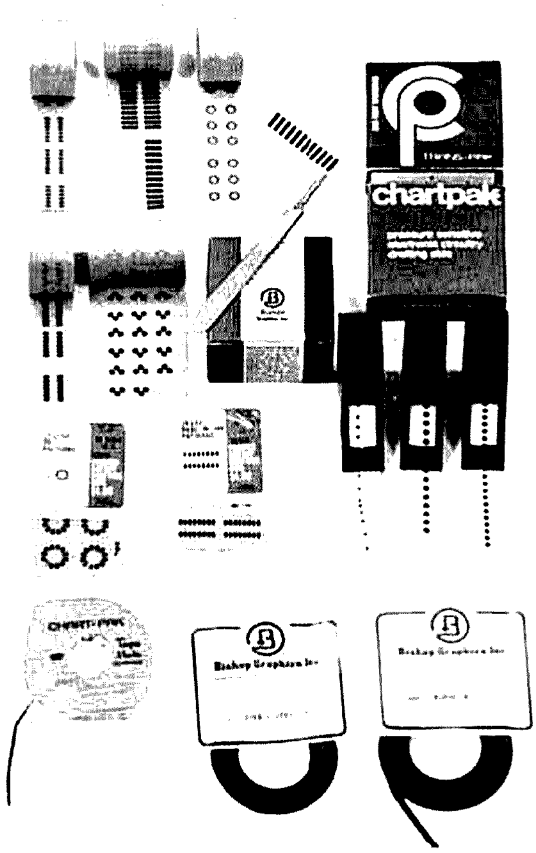


table 1. Drafting aids useful in laying out printed-circuit boards. Data is arranged by catalog number, number of images or feet of tape and price.

item	Bishop	Chartpak	Datak	GC Electronics
8 lead TO-5 (TO-99)	6026 250 $6.85	TPCL 11-4 250 $7.55	JD-71 1152 $6.50	21-103 20 $1.00
14 lead DIP (TO-116)	6014 250 $7.55	TPCL-12-1 100 $9.70	JD-68 432 $6.50	21-120 16 $1.00
.125" round pad	D207 500 $3.00	TPCC-72 250 $1.00	JD-124 4464 $6.50	21-023 125 $1.00
.187" round pad	D138 500 $2.00	TPCC-72 250 $1.00	JD-125 3120 $6.50	21-025 126 $1.00
.250" round pad	D109 500 $2.00	TPCC-77 250 $1.00	JD-127 1920 $6.50	21-025 126 $1.00
.020" tape	T201 (.020") 720 $.75	TP-1502 648 $.70	— — —	21-002 360 $.59
.040" tape	T201 (.040") $.75	TP-0401 $.70	— —	21-004 $.59
.080" tape	T201 (.080") 720 $.75	TP-0801 648 $.70	— — —	21-006 360 $.59

Bishop Graphics, Inc., 7300 Radford Avenue, North Hollywood, California 91605

Chartpak Rotex, One River Road, Leeds, Massachusetts 01053

Datak Corporation, 85 Highland Avenue, Passaic, New Jersey 07055

GC Electronics, 400 South Wyman Street, Rockford, Illinois 61101

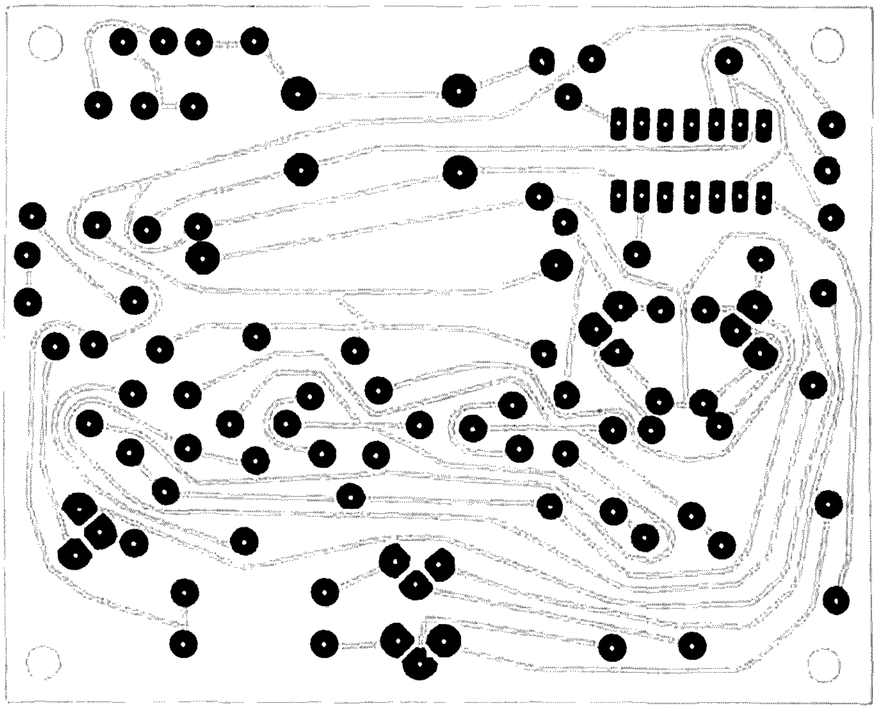


fig. 9. Place the round circuit pads, IC sockets and transistor layouts in the correct positions on the plastic sheet covering the reversed circuit drawing.

The board can be spun with an electric drill or small shaded pole motor mounted vertically on a small stand. The board is attached with a suction cup; a cardboard box should be used to catch the spin-off.

drying

The drying step is important because the photoresist must be dry to function properly but it must not be overheated. Shipley AZ-111 photoresist can be dried in an oven at 150° F (70° C) for five minutes. Kodak KPR-2 photoresist requires 20 minutes at 170° F (80° C).

Both types can be dried under a *red-filter* type heatlamp at a distance of about 9 inches for 5 to 10 minutes, using the gentle breeze from a fan to prevent the board from getting too hot. The photoresist can also be dried with a heat

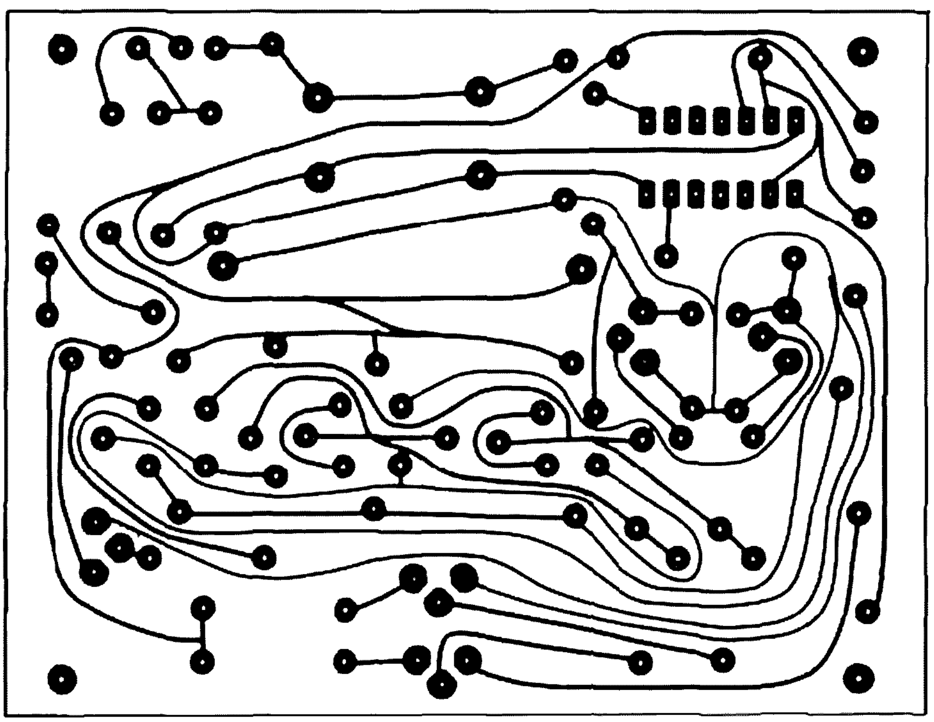


fig. 10. Connect the circuit pads with narrow black drafting tape. This is your completed printed-circuit layout transparency.

lamp while the board is being spun. With high-density boards make an effort to keep dust and lint off the work to prevent short circuits on the finished board.

exposure

The photoresist-coated board is placed in contact with the transparency and exposed to a source of ultraviolet light. The tape side goes on top unless you forgot to turn the paper over in the drafting step. To insure that the transparency and the copper-clad board are in close contact it is advisable to build a frame like that shown in fig. 12. If you don't use a frame simply place the copper-clad board on a flat surface, center the transparency, and place a sheet of glass over the combination.

A sunlamp at a distance of about one foot is used to expose the resist.* A fan should be used to keep the photoresist from getting too hot. The exposure time

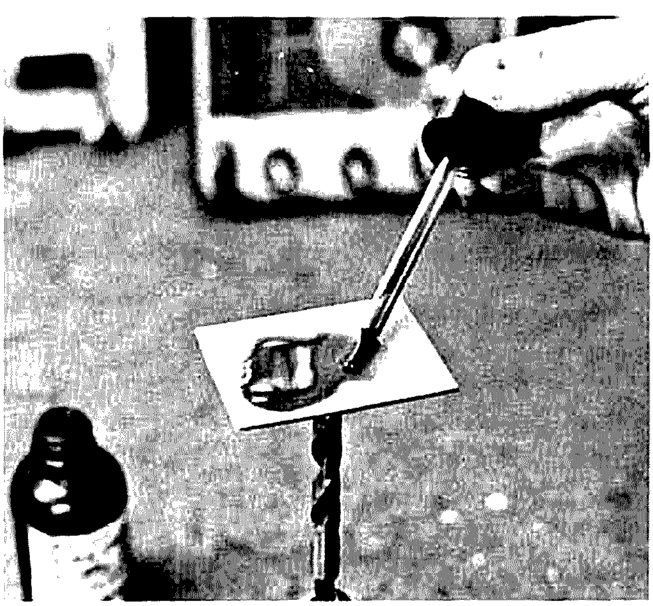


fig. 11. A thin, even coatf photoresist is applied to the board by spinning it with a drill motor.

for AZ-111 with the setup shown in fig. 13 is 15 minutes.

If you use a different ultraviolet source, a different lamp-to-board distance, or a different photoresist, you will have to determine exposure times with test strips. To do this you coat a piece of scrap board with photoresist, expose it, and with a piece of cardboard, cover or uncover a new section of board every few minutes. Since glass absorbs ultraviolet light it is necessary to place a sheet of the

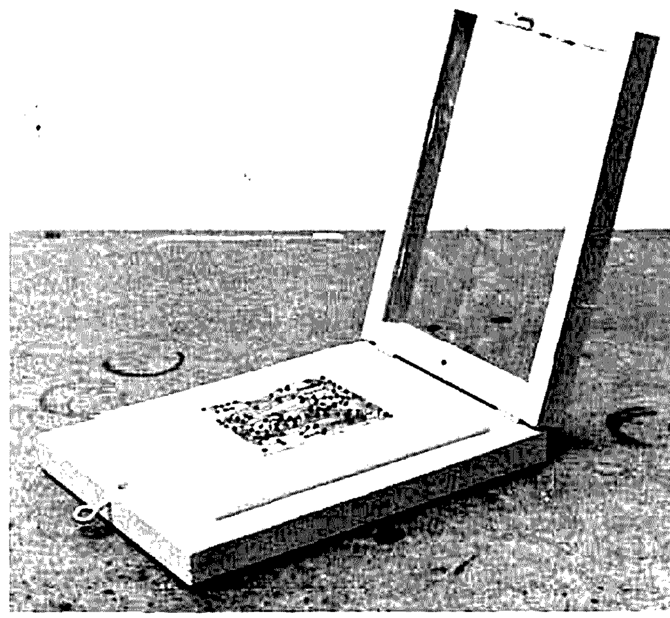


fig. 12. The circuit transparency and sensitized board are held in close contact with a home-made contact-printing frame.

same type glass between the lamp and the test strip. With a timed test strip you can quickly determine the proper exposure time for your combination of photoresist, sunlamp and lamp-to-board distance.

When choosing an ultraviolet source remember that the more the lamp resembles a point source the greater the degree of resolution. It is for this reason that fluorescent ultraviolet lamps are not completely satisfactory for highly detailed integrated-circuit work.

Room lighting is not critical for Shipley AZ-111. The dried boards can withstand up to two hours of ambient fluorescent lighting and considerably longer exposure to incandescent lighting. Other photoresists such as Kodak KPR-2 and Dynachem photoresist[2] are less tolerant in this respect due to their greater sensitivity.

*The sun is an excellent substitute.

developing

Shipley AZ-111 is developed by placing the exposed board in the developer at room temperature for about five minutes with occasional agitation. If the areas where the photoresist was intended to dissolve turn purple, your exposure was not long enough or your photoresist is too old.

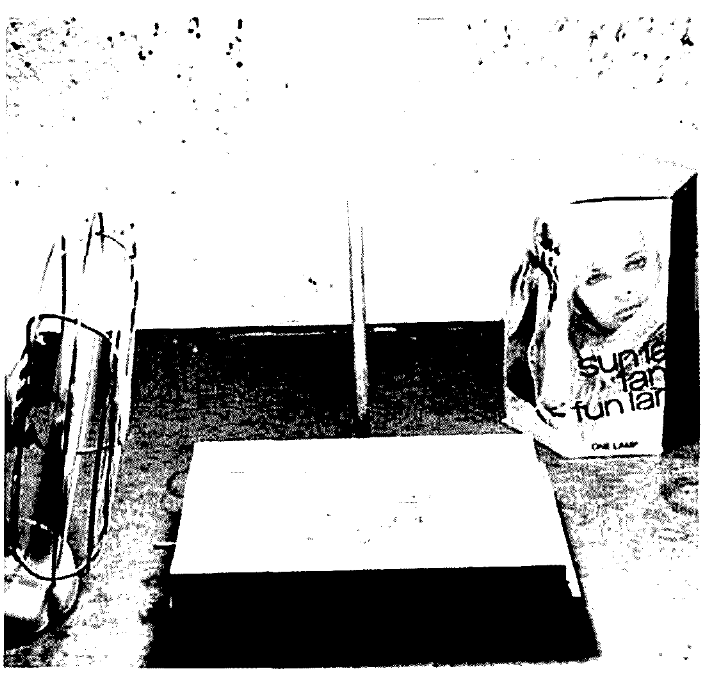


fig. 13. Circuit board is exposed to ultraviolet lamp for about 15 minutes. Fan keeps the photoresist from getting too hot.

The developer is made by dissolving 2½ teaspoons of household lye in one quart of water (15g/liter).* Use care when working with caustic solutions. After the board is developed, rinse it with cold water; it can now be etched.

Kodak KPR-2 is developed in trichloroethylene, a common dry-cleaning solvent, for about three minutes; work in a well-ventilated room. After developing, dry the board under a heatlamp for a few minutes. Do not touch the photoresist until it is completely dry (it is very soft while wet).

If you used a *negative* transparency with KPR-2 the board is ready to be etched; if you used a *positive* transparency the areas of copper you want to be etched away are covered with resist and those to remain are bare. Since the circuit would obviously not work if the board were etched at this point it is necessary to reverse the pattern on the board. This is done by immersing the board in a commercial electroless tin-plating solution. (See later section on plating solutions.)

When this is complete the resist is stripped off with acetone and a paper towel. The areas which are to remain after etching are plated with tin while the areas to be removed are bare copper. The board is now etched in chromic acid since it rapidly dissolves copper but does not attack tin.

etching

Some of the methods used to etch printed-circuit boards, according to speed and freedom from undercutting, are

1. Spray etching
2. Bubble etching
3. Splash etching
4. Immersion with continuous agitation or stirring
5. Simple immersion with occasional agitation

Bubble etching is the most highly recommended method for amateur use as it is very fast and does not require complex equipment. In ferric chloride at room temperature a two-ounce copper-clad board requires 60 minutes to etch by method 5, 15 minutes by method 4, and only 5 minutes with bubble etching.

A bubble etcher can consist of a large ceramic aquarium aeration stone immersed in the etchant and supplied with a source of air such as that provided by a heavy-duty aquarium pump. The circuit board is placed slightly below or on the surface of the etchant and inclined so there is a good flow of etchant along the entire under-surface of the board (figs. 14 and 15). Do the etching in a well ventilated room with a damp cloth over the etching tank to catch the spray. The

*The developer is deactivated upon long exposure to the air.

spray from the chromic acid etchant is especially hazardous.

etchants

A description of most of the etchants suitable for amateur use is provided by fig. 17. The vertical axis is the etching rate relative to cupric chloride; the horizontal axis is the degree to which the solution has been depleted by previous etching. It can be seen that ammonium persulfate is the fastest etchant shown and that the etching rate is increased considerably by the addition of a trace of mercuric chloride.

This graph also shows that the addition of hydrochloric acid to ferric chloride results in a decreased etching rate but allows the solution to be regenerated by simple overnight aeration. Ammonium persulfate often can be purchased from local electronics suppliers or from mail order houses; ferric chloride can be purchased in liquid form from local electronics suppliers. However, all etchants can be made much less expensively than they can be bought; the ingredients can usually be obtained from chemical supply houses in the larger cities. Buy the most inexpensive grade available, usually called *technical.*

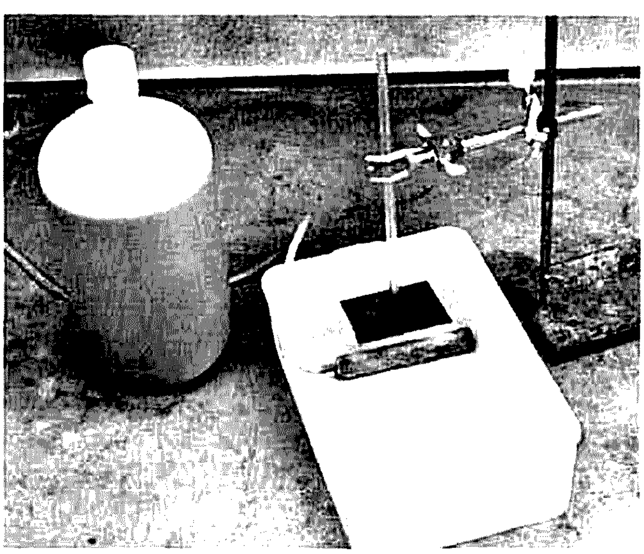


fig. 14. Circuit board is etched in large non-corrosive container. Oblong object under the circuit board is a ceramic aquarium aeration stone.

All etchants are toxic and should not be allowed to contact the skin or eyes. Remember that you must add acid to water, not the other way around. It goes without saying that you must keep all chemicals out of the reach of children.

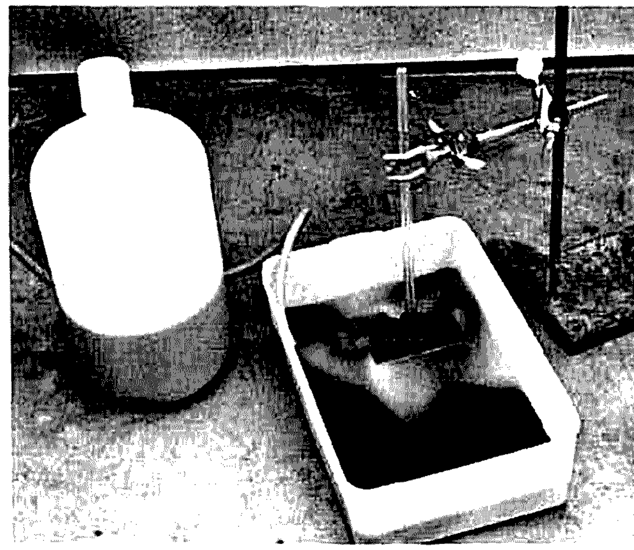


fig. 15. With the bubble etching technique the circuit board is completely etched in about 5 minutes.

Cupric chloride is my favorite etchant. Although it is slower than the others it is less expensive and is easily regenerated. Bubbling air gently through the solution for a few hours or overnight and an occasional addition of water and hydrochloric acid results in complete regeneration of the solution. Thus, after etching a few boards you end up with more etchant than you started with.

You will note that after etching a board the cupric chloride solution is quite dark. Aeration is complete when the clear yellow-green color reappears. Add 150 milliliters of concentrated hydrochloric acid and 350 milliliters of water to the etchant after using 1 liter to etch 400 to 500 square centimeters (approximately 70 square inches) of 2-ounce copper-clad circuit board.

plating

To obtain maximum corrosion resistance and solderability the photoresist should be stripped off with acetone and a

paper towel, the copper areas cleaned gently with a cleanser, and the circuit board immersed in an electroless tin-plating bath; nickel and gold are possible substitutes for the tin. You may want to

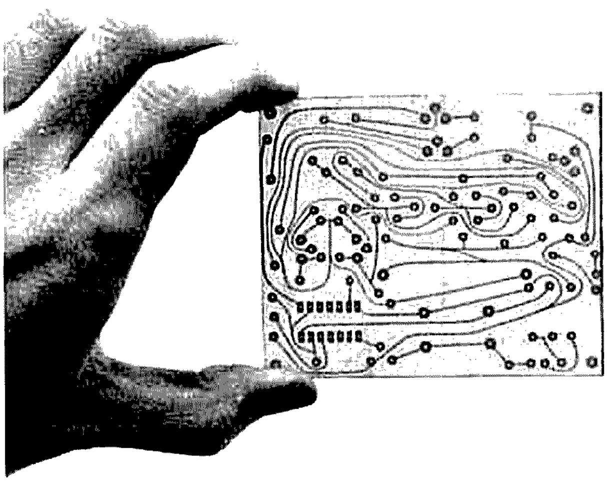


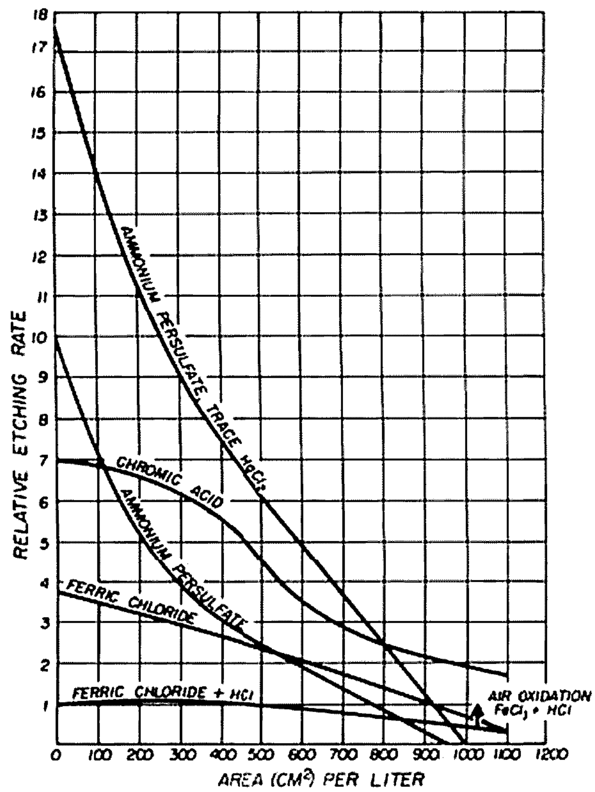
fig. 16. Completed printed-circuit board.

plate edge-connecting circuit boards with gold because of its high corrosion resistance. Unfortunately, Shipley gold- and nickel-plating solutions are presently available only in large quantities.*

Note that if you use a positive-transparency — positive-resist or negative-transparency — negative-resist combination you may plate before resist application *or* after etching. If you don't have confidence in your abilities as a chemist and choose not to plate, leave the photoresist on since it affords good corrosion protection and makes soldering only slightly more difficult.

electroless plating solutions

Electroless plating solutions are available which will plate a variety of metals at reasonable rates onto copper and certain other metal surfaces, or onto pretreated nonmetallic surfaces. No dc power is necessary as the solutions contain a reducing agent which chemically reduces the metal ions to the free metal at the solution-substrate interface. The catch to an otherwise beautiful story is that there does not appear to be an electroless tin bath which is stable, plates at room temperature, and produces a shiny, thick and pore-free deposit.



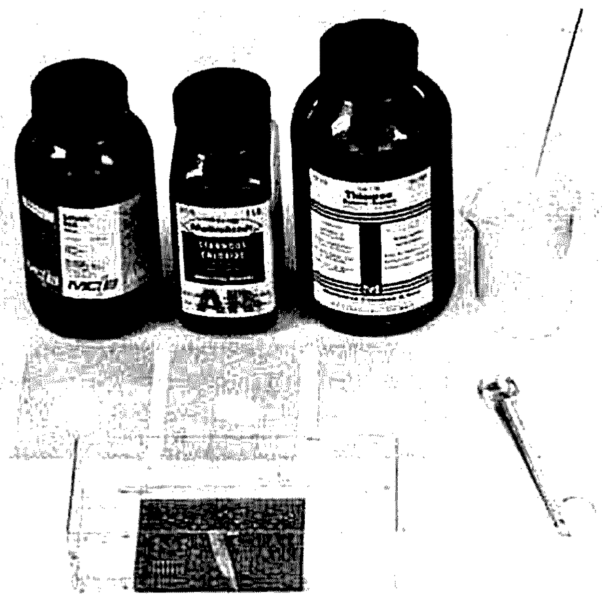
Ammonium persulfate:	240 grams ammonium persulfate per liter (plus 10 milligrams mercuric chloride)
Chromic acid:	200 grams chromium trioxide and 200 milliliters concentrated sulfuric acid per liter
Ferric chloride:	600 grams ferric chloride per liter
Ferric chloride plus hydrochloric acid:	600 grams ferric chloride plus 300 milliliters concentrated hydrochloric acid per liter
Cupric chloride:	100 grams cupric chloride plus 300 milliliters concentrated hydrochloric acid per liter

fig. 17. Etching rates of various etchants vs depletion of etchant, square centimeters of 2-ounce copper-clad board etched per liter of etchant.

A common electroless tin-plating bath which will produce a fairly shiny, if thin and porous, tin plate at room temperature consists of ½ gram (¼ teaspoon) stannous chloride, 2 grams (½ teaspoon)

*Gold plating, Shipley EL-221 gold, $24.50 per gallon. Nickel plating, Shipley NL-63, $19.00 per gallon. Shipley Company, Inc., 2300 Washington Street, Newton, Massachusetts 02162.

thiourea, and 3 grams (½ teaspoon) sulfamic acid dissolved in 100 milliliters (½ cup) of water. Five milligrams of either alizarin or alizarin red S may be added as a brightener if desired.



Materials for an electroless tin-plating bath.

This bath is unstable and only enough should be prepared at one time to cover the surface of the work to be plated. Plating time is 10 to 20 minutes. The bath does not produce deposits thick enough to withstand chromic acid and thus is not suitable for positive-mask — negative-resist — tinplate applications. For thick corrosion-resistant tinplate you must go to electroplating or high-temperature electroless tin baths. An example of the latter is Shipley's LT-27 tin (evaluation kit, $8.50). This bath is operated at 150-180° F with plating times of five minutes; bath makeup requires concentrated hydrochloric acid.

ground-plane circuit boards

In those instances where a good ground is necessary or where feedback may be a problem it is recommended that you use a ground-plane board. This can be done quite simply by using a two-sided copper-clad laminate and not etching one side. To prevent component leads from shorting to the ground plate it is necessary to make a guard ring around each hole. This can be done either by routing with a drill or by photoresist techniques.

When working with uhf circuits it may be advantageous to combine ground-plane printed circuits with a recently introduced form of point-to-point wiring. The ground lead of a component is soldered directly to the ground plane; the other leads are connected either point-to-point, to stand-offs, or to islands of copper isolated from the ground plane. Pieces of circuit-board may be soldered edge-on to the ground plane to serve as shields or as mounts for coil forms and variable capacitors.

materials

There is a considerable amount of high quality double-sided G-10 copper-clad board on the surplus market. Don't buy new printed-circuit board unless you are willing to pay about three times the cost of surplus material.

Kodak photoresists are available from local Kodak distributors ($14.30 per quart for KPR 2). Kodak Autopositive photoresist type 3 (KAR 3) is a positive-working photoresist comparable to Shipley AZ-111. (KAR 3 is $24.55 per quart; developer is $3.95 for 2 gallons.)

Shipley AZ-111 photoresist can be obtained from S&H Electronics* in quantities of 1 fluid ounce for $2, or direct from Shipley for $22 per quart. Large radio clubs might buy a quart and distribute the material to the membership. The shelf life of AZ-111 is one year at room temperature, two years if refrigerated. (Do not store in your home refrigerator if you have children who might mistake the photoresist for steak sauce.)

*S&H Electronics, Post Office Box 286, Corvallis, Oregon 97330.

references

1. "Printed-Circuit Handbook," GC Electronics, available from GC Distributors.
2. Thomas R. Rosica, "Making PC Boards from the Printed Page," *Popular Electronics,* October, 1968, page 69.
3. "Photofabrication of Printed Circuits," Eastman Kodak publication P-125, $1.00 from Department 454, Eastman Kodak Company, Rochester, New York 14650.

ham radio

top-loaded vertical

for 80 meters

Ideas and construction notes for a popular version of the 1/4-wave vertical antenna using a capacitive hat for top loading

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The purpose of this article is to pass along my experiences and some ideas on the evolution and construction of an 80-meter top-loaded vertical antenna. To keep things in chronological order, I'll start with the simple vertical wire I first put up. It consisted of a 60-foot length of no. 12 wire suspended from an insulator attached to the limb of a tree. The top of this antenna was 65 feet above ground.

ground losses

All antenna literature states that a good ground system is necessary for a base-excited ¼-wave vertical. Therefore I laid out 24 radials around the antenna base. Each radial was no. 12 wire 66 feet long. But is this a good ground system for 80 meters? I thought so, and friend W2LV agreed it was the best I could do under my circumstances. However, W2LV kept mentioning ground losses and finally provided the information in **table 1**. This data doesn't exactly fit my case, but it is representative and based on a vertical radiator of 0.2 wavelength. The data goes back to the 1930s when broadcasters were interested in standards for antennas with vertical propagation angles.

The interesting thing about the data is that if the radials are too short, say 0.15 wavelength, they don't do much good.

impedance matching

Using a GR 916A rf bridge, I made measurements between the antenna and ground system. Antenna characteristics are shown in **fig. 1.** A pure resistance of 57 ohms appears at 3.65 MHz. I wanted the antenna to resonate at 3.8 MHz, but at this frequency the impedance was 63 + j36 ohms, which is equivalent to an swr of 2. This indicated that some kind of matching arrangement was necessary.

The series-resonant circuit shown in **fig. 2A** was constructed and installed in a plywood box, then mounted on a short post at the bottom of the antenna wire. The 50-ohm tap point for the coax cable was determined using a homemade bridge.[1] A point on the coil producing a zero indication could not be found, so the arrangement of **fig. 2B** was used. The tap and capacitors were carefully adjusted to give a null. (Incidentally, I found from experience that it's prudent to check an impedance bridge with a carbon composition resistor to make sure everything is working properly.)

The transmission line was connected, and the transmitter was fired up. An swr of 1.2 (at 3.8 MHz) was obtained at the transmitter end with an swr bridge.

fig. 1. Measurements of the initial 80-meter ¼-wave vertical suspended from a tree limb. Desired resonant frequency was 3.8 MHz.

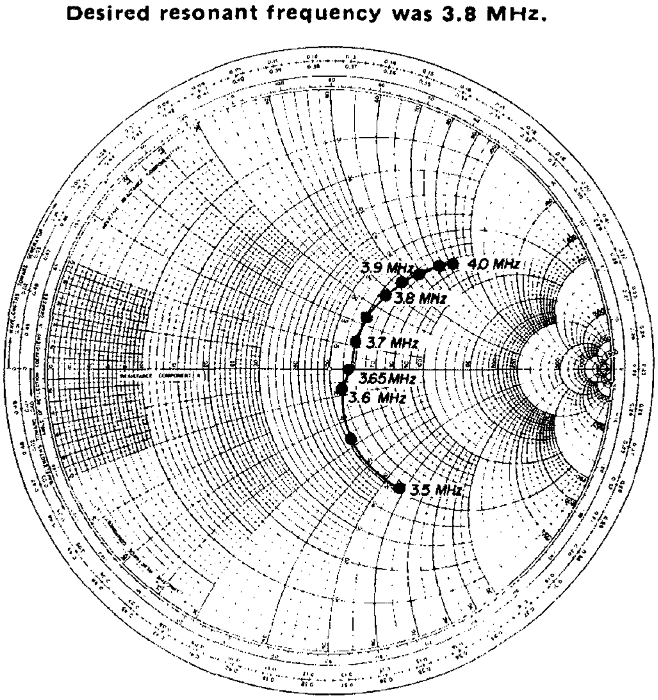

table 1. Ground loss in radials for vertical antennas.

no. radials	radial wavelength	percent ground loss
30	0.4	23
30	0.15	41
60	0.4	10
60	0.15	38
120	0.4	2
120	0.15	35

Results were interesting. Close-in signals (200-300 miles) were weak on the vertical compared to a horizontal antenna. Signals further away were about equal in amplitude on either antenna. DX signals (Europe, Hawaii) were stronger on

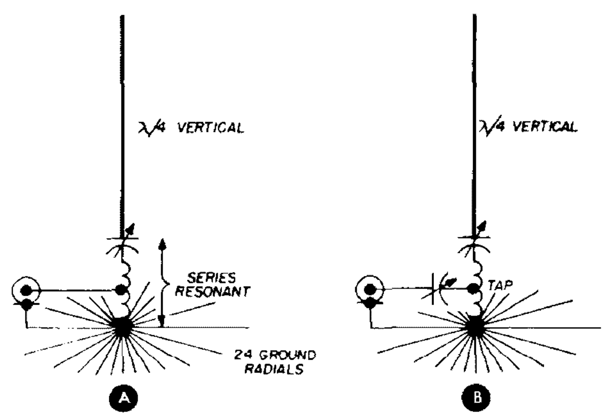

fig. 2. Tuned circuits for resonating the initial version of the vertical. Circuit at B produced an swr of 1.2 at the desired operating frequency.

the vertical. I worked 23 countries on 80-meter phone in the 1969 ARRL DX test with the vertical.

top loading

Flushed with success, my next objective was to improve the efficiency of the vertical by moving the current loop away from ground. Top loading seemed to be the answer.

After reading the early articles [2, 3, 4] the installation shown in **fig.** 3 evolved. Although top loading can be accomplished with a ball, cylinder, or disk, the latter seemed the easiest to build. A chart[2] of capacitance vs. disk diameter showed that a disk six feet in diameter would give a capacitance of 61 pF. The only reason I picked six feet was that it was a nice

number and I felt the bigger the better. I had reason to regret this decision, which I'll now explain. You'll notice from **fig. 3** that a pulley arrangement is used to raise and lower the antenna. Such a method is convenient and saves climbing during tuning adjustments. However, be sure the vertical path is free of obstructions for the whole diameter of the disk! I had a fearful journey up the tree cutting down branches which criss-crossed the six foot imaginary cylinder through which the disk has to pass. There were times when I wished that a 1-foot-diameter disk had been used.

capacitive hat

The disk was made by using two six-foot lengths of 1 x 2 inch wood in the form of a cross. A circle of ¼-inch copper tubing was supported on standoff insulators at the ends of the cross; and 24 no. 12 wires were connected from the outside to the center of the disk (**fig. 4**).

loading inductance

Knowledge of the disk capacitance is necessary to obtain a ball-park figure for the loading-coil inductance. An inductance of 29 μH was required to resonate the system at 3.8 MHz with the 61-pF disk.

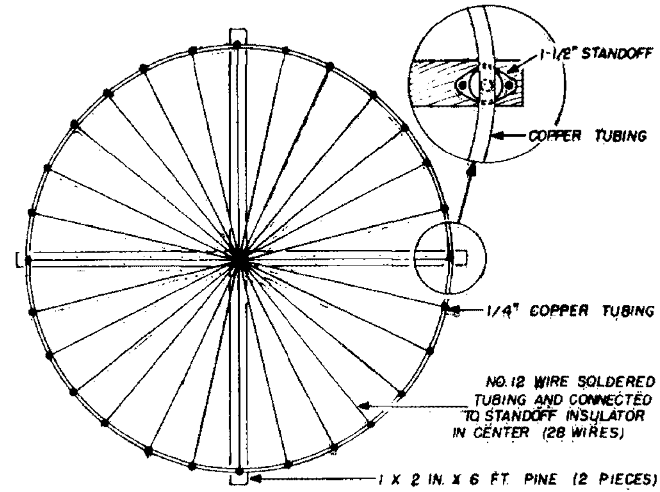

fig. 4. Construction details of the capacitive hat. The copper tubing is flattened and attached to a standoff insulator as shown in the inset.

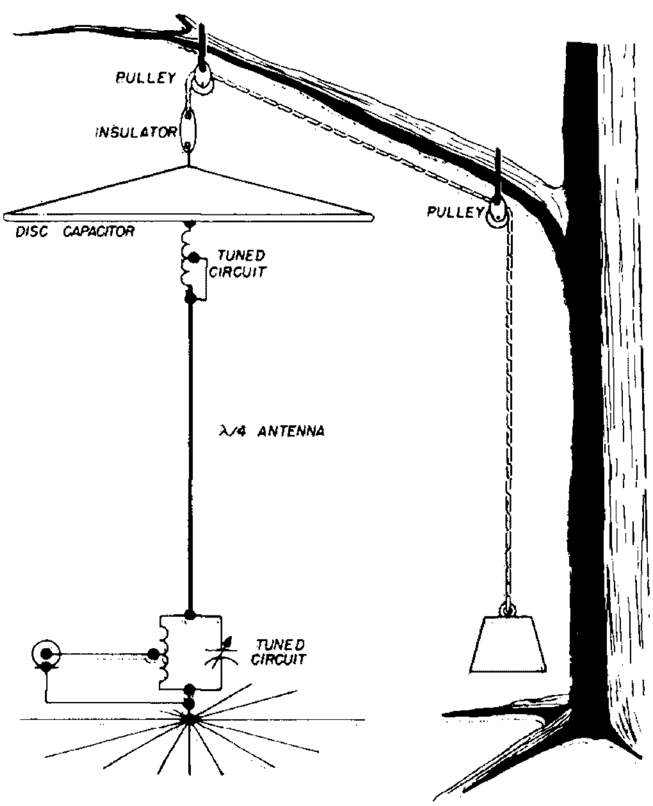

fig. 3. The antenna with a capacitive hat for top loading. The disk was six feet in diameter, but a smaller disk could have been used with a larger tuning inductance.

The coil was from some surplus equipment. It has a 3-inch-diameter ceramic form with 33½ turns of no. 12 wire spaced the diameter of the wire, with an overall inductance of 44 μH. This coil, with provisions for shorting out turns, was mounted as shown in **fig. 3** and the whole thing hoisted in the air.

Now, how do you tune the coil to bring the antenna into resonance at the desired operating frequency? The old QST articles stated that the field strength of the antenna should be monitored at a distance greater than 300 feet. A point is reached when decreasing the coil inductance causes the field strength to drop off. You then go back a few turns and there you are.

Such a method works but it doesn't seem very scientific. Since I had an impedance bridge available I decided to do it properly.

Unfortunately the impedance bridge

measured only to about 1200 ohms, and the base of a properly tuned top-loaded vertical is probably several thousand ohms. It was necessary to have some means of using the bridge while adjusting the coil turns to produce a pure resistance at the base of the antenna, which would be in the range of the rf bridge. If

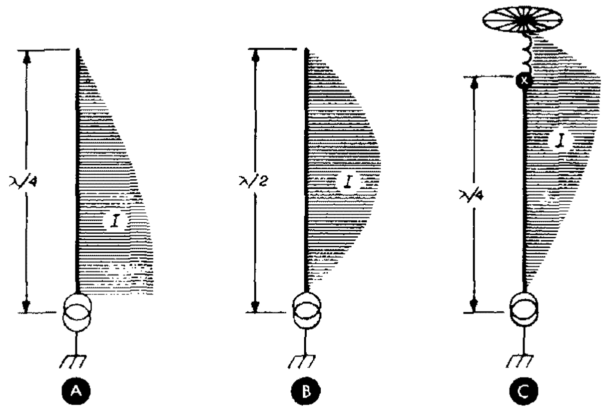

fig. 5. Current distribution on three vertical antennas. The tuned circuit at C simulates ¼ wavelength.

a pure resistance is obtained at the antenna base, the antenna is in resonance.

measurements

Let's pause a minute to look at the problem. **Fig. 5** shows several conditions of current distribution. A is that of the quarter-wave vertical, B a half wave vertical, and C a top-loaded quarter-wave vertical to simulate a half wavelength. If we could put our measuring equipment at the top of the antenna, a low-impedance point would be available at point X; not very practical.

W2KXD pointed out that the same objective could be accomplished at ground level by using a quarter-wave extension as shown in **fig. 6.** The quarter-wave coax will translate the high impedance at the antenna end to a very low impedance at the bridge end of the coax. The bridge measurement will show a pure resistance when the antenna is at resonance. I believe this method of tuning the coil of the top-loaded vertical is novel.

Don't let the mention of a General Radio rf bridge scare you off if you don't have one available. The rf bridge described in Reference 5 will work very well. The quarter-wave coax length was determined by measuring its input with the output open circuited, and cutting until a pure resistance input was obtained. Before connecting the antenna, I took the precaution of inserting a 50-ohm carbon composition resistor across the bridge. It read 49.5 ohms, so I knew my setup was correct. With the total coil in use, a reading was taken at 3.5 and 4.0 MHz with the following results:

3.5 MHz Z = 1.0 + j0 ohms

4.0 MHz Z = 1.5 + j 12.1 ohms

The pure resistance at 3.5 MHz indicated that the antenna was resonating at that frequency, but 3.8 MHz was desired.

The antenna was lowered and eleven turns shorted out on the coil. After pulling the antenna back into place, the measurements then became:

freq (MHz)	impedance (ohms)
3.5	1 – j7.1
3.6	1.3 – j4.2
3.7	1.3 – j2.7
3.8	1.7 – j0
3.9	1.7 + j2.7
4.0	1.5 + j5.0

Resonance is indicated at 3.8 MHz, just where I wanted it. To further test the procedure, 23 turns were shorted out with the following results:

freq (MHz)	impedance (ohms)
3.5	2.5 – j11.4
3.6	2.0 – j9.7
3.7	2.2 – j4.0
3.8	2.3 – j2.6
3.9	2.2 – j2.0
4.00	2.2 – j1.2

It can be seen that resonance is above 4.0 MHz. Therefore the 11-turn tap point was correct.

Since there was now a high impedance at the base of the antenna, a parallel-tuned circuit was used as shown in fig. 7 and adjusted as previously described.

After reconnecting everything, the transmitter was fired up. The swr bridge at the transmitter end showed the following readings:

frequency (MHz)	swr
3.5	almost full scale
3.6	3.3
3.7	1.4
3.8	1.2
3.9	1.5
4.0	3.0

A check on the receiver showed that locals were way down compared to those on the half-wave horizontal, and equal signal strength between the two antennas appeared to be at about 1000 miles. The vertical was used on 80 meters during the 1969 phone CQ DX contest with good results. I had the intuitive feeling, however, that I could work DX stations just as well with the half-wave horizontal, because received signal strengths seemed to be the same. It was not working the way I had hoped it would, and I wanted to do some more testing.

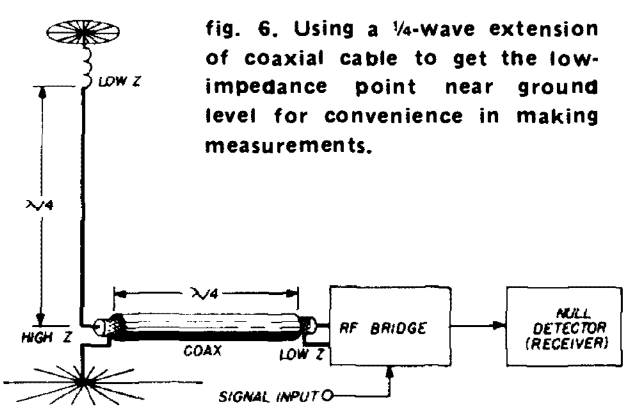

fig. 6. Using a ¼-wave extension of coaxial cable to get the low-impedance point near ground level for convenience in making measurements.

current distribution

One nice thing about having a project and friends is the interesting discussions one can get into. W2LV suggested that I get rid of the coil and just use the disk, his feeling being that the coil is a loss unless it's wound with ¼-inch copper tubing, silver plated.

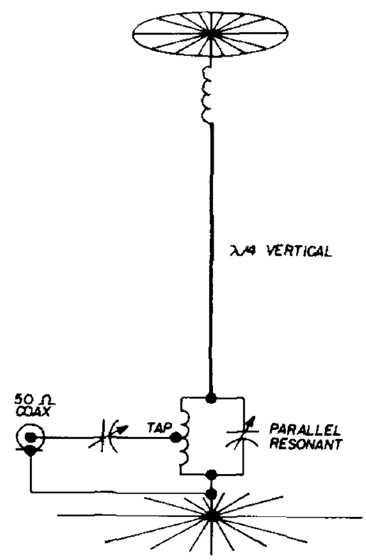

fig. 7. Parallel-resonant circuit for tuning the final configuration.

W2LL suggested that I solder a bunch of Christmas tree lights along the antenna and see if the current distribution was correct, then I could leave them up for the holiday season.

One never knows quite when to believe W2LL, but his suggestion fired my imagination. A dozen 2-volt, 60-mA bulbs were purchased and spaced every five feet along the 60-foot wire. There was some discussion on how far to tap the bulbs across the wire, but little objection to my suggestion of 5 inches. Having the antenna on the pulley made it easy to lower it, disconnect it, roll it up, and take it indoors to work on. The next evening, with the bulbs in place, the antenna was reconnected and hauled into position. It took 1 kW of power to light the lamps to proper brilliancy.

final adjustments

I looked at the lights from all angles and several distances, trying to figure out a way to record their brilliance quantitatively. How do I get at a point equally distant from each light? Can I photograph them?

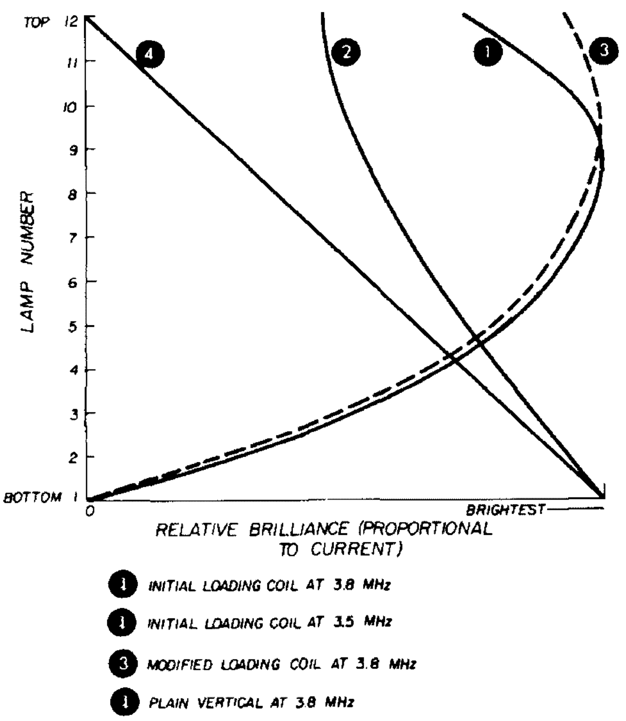

fig. 8. Results of tests using small lamps to determine current distribution.

The final solution was to pick out the brightest lamps, and those above and below that point were grouped as to equal brilliance. Very crude of course; but if you are there, you know what's happening. A plot of what I saw of the current distribution on the top-loaded ¼-wave vertical antenna is shown in fig. 8 for 3.8 MHz. The lamps at 1 show a current loop about ¾ of the way to the top of the antenna. The transmitting frequency was increased to 4.0 MHz with results approximately the same. The transmitter frequency was then reduced to 3.5 MHz, and the lamps assumed a relative brilliance as shown by 2.

The next step was to lower the antenna, short out the total loading coil, and pull up the antenna. With the transmitting frequency returned to 3.8 MHz, the relative current distribution appeared to be as at 2.

The light presentations for the different displays agree with theory except for the 1 readings; the maximum brilliance should be at the top.

Down came the antenna, and coil removed. The coil Q measured 290 and had an inductance of 23 μH. (The same coil with the shorted turns unshorted had a Q of 400 and an inductance of 44 μH.)

Turns were removed from the coil to return its inductance to 23 μH. This not only improved the Q of the coil by eliminating the shorted turns, but improved its form factor as well. The final Q of the coil at 3.8 MHz was 480; a 165% improvement over the original value.

The coil was reinstalled in the antenna and the lights inspected. The dashed line, 3, of fig. 8 shows the improvement in the current distribution. I have no doubt that W2LV is correct in suggesting the use of ¼-inch copper tubing for the coil – or better yet eliminating it – but the improvement one way or the other seemed marginal.

conclusions

I've attempted to pass along my experiences and some ideas on the evolution and design of an 80-meter top-loaded vertical antenna. You might say that my results were inconclusive since top loading gave no better results than base loading. However, top loading does get the current loop at the highest point of the antenna. If that's what you want, now you know how to do it.

references

1. H. O. Pattison, W2MYH, et al, "A Standing Wave Meter for Coaxial Lines," *QST,* July, 1947.
2. R. B. Dome, "Increased Radiating Efficiency for Short Antennas," *QST,* September, 1934.
3. T. M. Ferril, Jr., W1LJI, "Simple Vertical Antenna," *QST,* February, 1939.
4. Paul H. Lee, W3JM, "Vertical Antennas," *CQ,* June, 1968-May, 1969.
5. Henry S. Keen, W2CTK, "RF Impedance Bridge," *ham radio,* September, 1970.

ham radio

frequency synchronization for scatter-mode propagation

A frequency standard that can be used by vhf/uhf stations to precalibrate receiving systems

James A Buscemi, K2OVS, A1L, A Division of Cutler-Hammer, Inc., Deer Park, New York

Vhf and uhf stations attempting to communicate via the more exotic types of propagation (meteor scatter, moonbounce, tropo scatter, etc.) require close synchronization in two areas: time and frequency.

the frequency sync problem

Time synchronization is generally not a problem. WWV, CHU, and other sources provide a time reference accurate enough for most communications. Accuracy of the receiving and transmitting frequencies is quite another matter, however.

Most moonbounce and meteor scatter work requires an extremely narrow i-f bandwidth (100-500 Hz) to enhance the effective signal-to-noise ratio of the receiving system. Setting this extremely narrow window accurately on the desired receiving frequency is difficult at best. At 432 MHz, 100 kHz represents a frequency accuracy of 2×10^{-5} percent! Most standard crystal calibrators or signal generators have inherent accuracies at least 100 times worse than this, not considering temperature variations. In addition, most receiving systems use at least one frequency conversion with a crystal-controlled local oscillator, resulting in another frequency error; namely, the accuracy and frequency stability of the crystal in the converter local oscillator. Multiple frequency conversion systems, commonly used above 420 MHz, are proportionately worse. Even a well-equipped vhf-uhf amateur station has the capability to calibrate only the station receiver, not the entire receiving system. As a result, most scatter-mode communication failures occur because one station or another is listening on the wrong frequency! This has happened on occasion to even the most proficient operators.

temperature compensation

One possibility that comes to mind is to temperature compensate all frequency-determining components in the system and perform a one-time calibration, assuming all components do not drift or age with time or temperature. Experience has shown this to be a poor assumption;

periodic recalibration of all test and measuring equipment is standard practice in industrial and military establishments. In addition, most amateurs don't have access to frequency counters or sophisticated lab equipment.

frequency standard

Some sort of periodic frequency-calibration method is therefore necessary sensitivity or antenna testing. A calibrator such as this is a worthwhile accessory and could, theoretically, be adjusted to extreme accuracy by comparison with an NBS traceable standard. However, individual construction and use of a similar calibrator would only introduce additional errors between stations (even commercial standard frequency counters have been known to drift out of calibration).

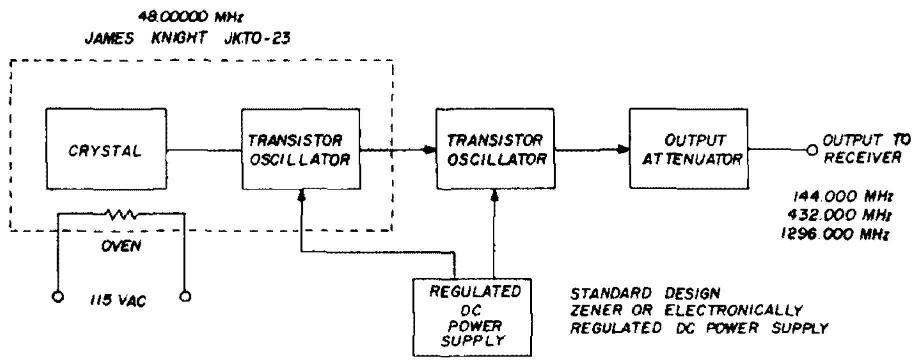

fig. 1. Suggested frequency standard for vhf/uhf moonbounce and tropo scatter work. Unit would be used by participating stations on a sharing basis to assure frequency synchronization before scheduled contacts.

for stations involved in moonbounce or meteor scatter work. A frequency-standard source is a necessity; a typical source is outlined in fig. 1. This frequency standard is crystal controlled, temperature stabilized and compensated,* and voltage regulated to provide extremely stable frequency markers on 144, 432 and 1296 MHz. A James Knight type JKTO-23 oscillator module at 48 MHz provides the basic frequency stability. The oscillator alone produces enough output to calibrate most receiving systems on these frequencies and allows calibration of the transmitting frequency by comparison on the station receiver. The addition of amplifier, multiplier and/or attenuation stages is optional, depending on individual needs and desired flexibility.

other uses

If calibrated on a power-output basis, the unit may also be used for system

*Readers wishing to build their own crystal ovens will find an interesting proportional temperature control scheme in reference 1. **editor.**

Therefore, since most moonbounce, meteor scatter, and tropo work is generally conducted on a schedule basis between a small number of stations whose operators are known to each other, it is proposed that *one particular frequency standard* be constructed and circulated via parcel post among these operators immediately before planned schedules. Each operator could then use the same standard to calibrate his receiving (and transmitting) equipment before the schedule, thus increasing his chances of success.

Even if the standard had some long-term frequency drift, the stations using it would still be on the same frequency, regardless of whether that frequency were 432.0001 or 431.9999 MHz.

I hope the foregoing proposal will be of value to fellows involved in vhf/uhf work and promote more successful contacts on the bands above 144 MHz.

reference

1. J. A. Koehler, VE5FP, "Proportional Temperature Control for Crystal Ovens," *ham radio*, January, 1970, p. 44.

ham radio

high power injection lasers

More on the characteristics and driving requirements of these remarkable little lasers

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With the continuing crowding of the radio spectrum it is likely, indeed probable, that forms of communication more esoteric than radio will eventually have a genuine usefulness to the amateur. Likely contenders are millimeter waves and optical links; the latter category is already within easy reach of today's amateur. It is the purpose of this article to review in detail some of the important characteristics of the diode laser and to provide some circuit information for driving high power units.

characteristics

Laser diodes are moderately expensive and relatively fragile electrically. Invented at almost the same time by three different research laboratories (GE, IBM and MIT) in the fall of 1962, the first devices were formed by the diffusion of zinc into n-doped wafers of gallium arsenide (Ga-As), the material still most commonly used for semiconductor lasers.

Mechanical polishing of GaAs ingots or cleavage along crystalline planes (the preferred technique for commercial devices) produces the two end mirrors necessary for formation of the lasing cavity. Lasing occurs when electrons which have been stimulated to a high energy level to cross the pn junction fall back to the ground state. In the process their stored energy is given off as a photon of light (though other emissions may occur on a less frequent basis).

The emission of photons may be stimulated by other photons, hence setting up a chain reaction situation and the condition of *stimulated emission.* It is the stimulated emission of radiation that results in the production of quasi-coherent light.

Since the laser diode has very high gain the length of its cavity may be very short — only 10 or 15 thousandths of an inch. This is fortunate as the threshold for lasing, which is dependent on the

diode's surface area, would be prohibitively high if a lengthy cavity were required for proper operation.

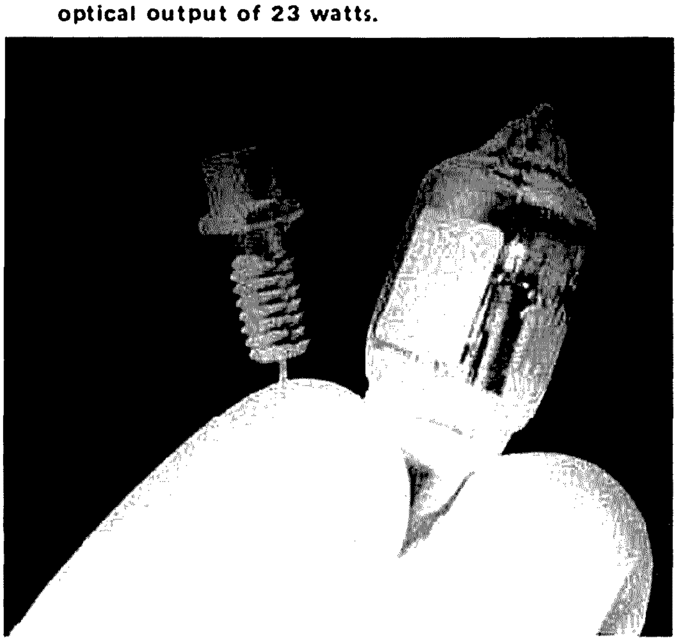
Though there are many possible configurations for experimental laser diodes commercial devices are either diffused or epitaxial. Epitaxial devices consist of two structures, homojunction and heterojunction. As the name might indicate the former is a straight GaAs epitaxial diode. The heterojunction, however, is formed with one side of the junction being composed of gallium aluminum arsenide (GaA1As) a material similar in characteristics to GaAs.

Since the index of refraction of GaA1As is higher than that of GaAs light radiating from the junction is subjected to a waveguiding effect with a resultant increase in laser efficiency. The magnitude of this increase in efficiency can be illustrated by comparing the lasing threshold for a diffused and a heterojunction laser, typically 100,000 amps/cm^2 and 10,000 amps/cm^2, respectively.

cw lasers

One device which is still in the laboratory is the double heterostructure laser.

Both injection lasers (left) and Krytron tubes (right) are quite small. The laser shown here is an RCA TA7699, a moderately high-power unit with typical optical output of 23 watts.



Invented at Bell Telephone Laboratories in the summer of 1970, this is the first semiconductor laser to operate continuously at room temperature. A significant lowering of the lasing threshold to 2,000 amps/cm^2 resulted in this achievement. As the name implies, the double heterostructure device has a layer of GaA1As on either side of a central GaAs region. The enhanced waveguiding effect of the two GaA1As layers helps to lower the lasing threshold.

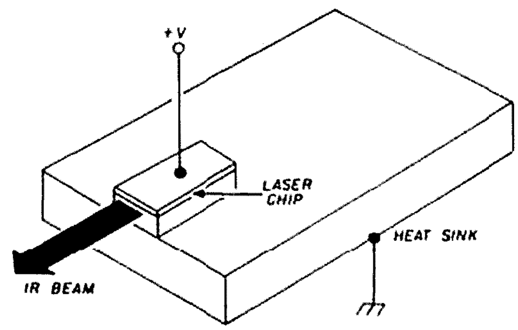


fig. 1. A typical injection laser is mounted as shown here. Infrared energy emerges from one of the parallel reflecting ends.

Though the new cw-at-room-temperature devices are not yet available commercially, their development is reported here because of their obvious application to voice communications. The ease with which these devices can be modulated greatly reduces the complexity and expense of both driving and receiver circuitry.

Like most semiconductors, laser diodes are sensitive to extremes of temperature and operating current. The lasing threshold is greatly dependent on temperature as shown in fig. 2. The sometimes subtle change from room to outdoor temperature can reduce the lasing threshold to the point where degradation can occur if the laser is operated at a current level suitable to the former condition.

Engineers at Holobeam, Inc. have designed a laser voice communicator with a feedback loop incorporating a thermistor in intimate contact with the laser to alleviate the problem. The thermistor

senses temperature changes and reduces laser current appropriately when temperature decreases.

high-current drivers

A previous article in *ham radio* provided information on driving lasers with a low to medium operating current.[1] However, lasers are currently manufactured by several companies which require 150 or more amperes for optimum operation. IBM has one single laser diode which radiates as much as 50 watts peak optical power from each end when driven by 300 amp pulses (100 nsec pulse width).

Though laser arrays capable of delivering up to 300 watts are available off the shelf from both RCA and Laser Diode Laboratories (and other companies as well) the ease with which a single diode laser may be collimated makes high power devices attractive candidates for long-range optical communication experiments.

It is difficult to design semiconductor drivers for producing currents greater than 100 amps. Hunt has mentioned SCRs that can switch currents of this magnitude;[2] and Brown, *et al* have presented an interesting driver consisting of a parallel arrangement of avalanche transistors which produces 200-amp pulses.[3]

An elegant solution to the problem of generating high current pulses is the miniature cold-cathode krytron (which should bring a chuckle from the anti-semiconductor club). Krytrons* achieve extremely rapid switching times when discharge is initiated between cathode and anode. The discharge, which propogates through a gas under low pressure, is normally initiated by a positive pulse on the grid.

The circuit shown in **fig. 3**, which is similar to one developed by Sullivan, is capable of delivering pulses with an amplitude of up to 2,000 amps![4] In operation, the eight 0.002-μF capacitors are charged through a resistor of several hundred thousand ohms until the krytron is triggered *on* by a positive pulse at the grid. The high voltage stored in the capacitors is then dumped through the krytron, laser and the 1-ohm resistor. The resistor permits the 50-nsec pulse to be monitored with a fast oscilloscope (since the resistor's value is 1 ohm, volts displayed on the scope equal current in amps through the laser).

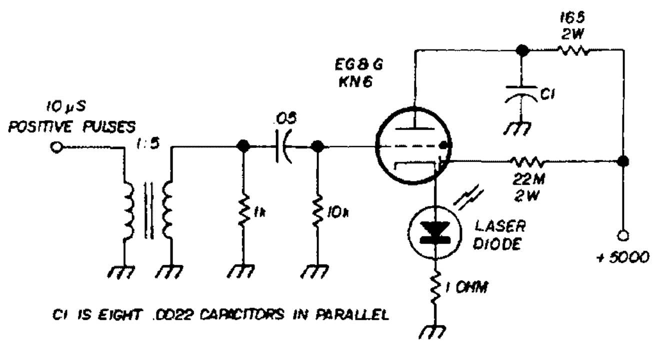


fig. 3. 2000-amp injection laser driver. Keep all leads carrying high current as short as possible. A unijunction-transistor oscillator will provide the positive-going input pulses.

Because of the inductance contributed by component lead lengths the pulse generator discharge section is actually an RCL circuit. To optimize the circuit's pulse shape L should be decreased as much as possible by reducing lead lengths to the absolute minimum. Otherwise the circuit's pulse will include negative components, be quite ragged in appearance and quite likely be lengthened.

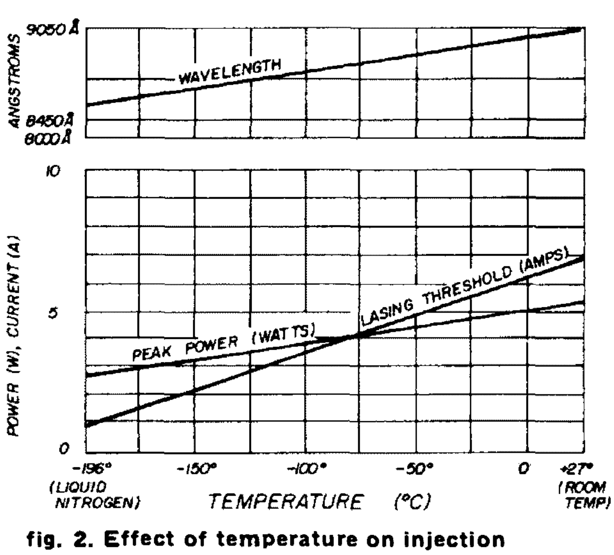


fig. 2. Effect of temperature on injection laser characteristics.

*Made by EG&G, 160 Brookline Avenue, Boston, Massachusetts 02215.

Of course, 2,000-amp pulses are overkill when it comes to driving lasers requiring from 100 to 300 amps. Peak current can be reduced to these lower values simply by lowering the charging voltage and/or slightly increasing the resistance of the current monitor. Peak current can be calculated from known component values, but because of the significant influence of undesirable strays, it is helpful to make actual current

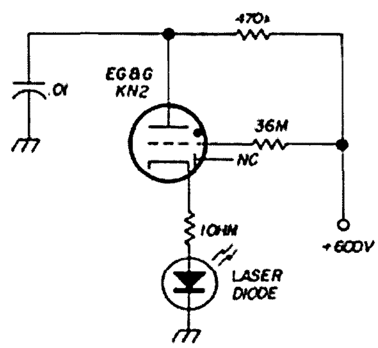


fig. 4. General-purpose injection laser driver. Keep all high current carrying leads as short as possible.

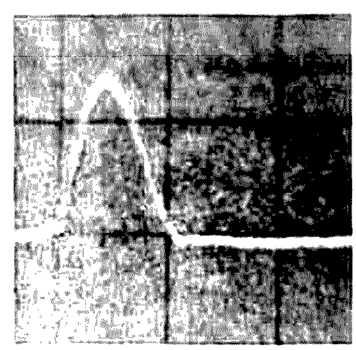
measurements with an oscilloscope using a conventional junction diode for a load. (To protect the laser diode it should be connected into a driving circuit *only* after peak current is known.)

The circuit in fig. 3 may be operated at pulse repetition rates of up to 1000 Hz. Although fig. 3 is preferred for applications requiring current adjustments, such as when a variety of lasers are to be used, the circuit in fig. 4 is perfectly adequate for experimental work. Adapted from a design by Koechner,[5] the circuit is a simple free-running relaxation oscillator. Positive bias on the grid pre-ionizes a portion of the gas in the krytron and prepares the tube for firing. When the voltage stored in the 0.01-μF capacitor becomes sufficiently high to fire the krytron, the capacitor discharges through the tube, the 1-ohm resistor and the laser diode. The 1-ohm resistor serves as a current-monitoring point as in the previous circuit.

With the values shown in fig. 4 the circuit will produce a 50 nsec, 75 amp pulse at a repetition rate of about 200 pulses per second. If lead lengths between all components carrying the current pulse are kept short very clean pulses will be obtained. The circuit is ideal for driving moderately high power single laser diodes; suitable alterations of appropriate component values will permit very high power single lasers, such as the IBM series and the RCA TA7705 and TA7787, to be driven as well. Modifications are easily accomplished with the help of a fast oscilloscope. Serious experimenters would be well advised to consult Koechner's very complete paper for additional help; his theoretical treatment is excellent.

power supplies

I have had good results operating both circuits with a miniature dc-dc converter. There are numerous ways of designing and building dc-dc converters; with the



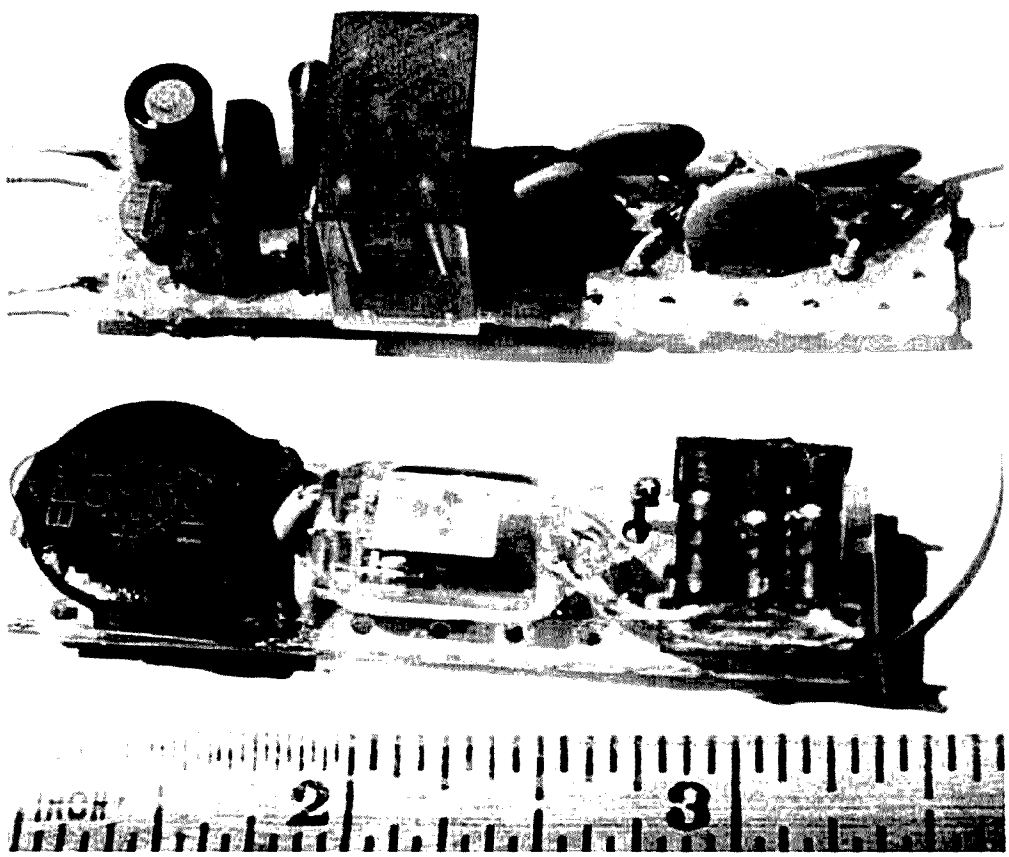
The circuit of fig. 4 yields an excellent pulse. Calibration is 100 nsec per horizontal division and 50 amps per vertical division.

numerous references available there is no need to describe a particular circuit here. Suffice it to say that a transistorized blocking oscillator with an appropriate step-up transformer followed, perhaps, by a voltage multiplier will permit a krytron laser driver to operate from a small 3-volt battery.

High-voltage photoflash batteries may be used to obtain the several hundred plus volts necessary to operate a laser driver, but the miniature size of the dc-dc converter makes it a very attractive power source, particularly for field experiments.

Additionally, the dc-dc converter is a one-time investment in contrast to high-voltage batteries. Regardless of the power source you choose be sure to follow all precautions that normally apply to high voltages. Size is deceiving — batteries and miniature dc-dc converters are capable of delivering healthy shocks.

With an impressive background of military contracts in the semiconductor laser business RCA offers, at present, the best buys for the experimenter in high-power single laser diodes. For low- to moderate-power lasers, try any of the companies except IBM. By the way, don't hesitate to write to the manufacturers for more information; most of them will gladly supply very good data sheets and price and delivery schedules.



Injection laser drivers and power supplies can be easily miniaturized. The dc-dc converter (top) can supply up to 600 volts and will operate the laser driver (below) when powered by a 3-volt battery. The cube of 0.1-ohm resistors provides an especially low-inductance current monitor. These units were developed by the author for use in a small rocket.

parts suppliers

Only a few companies manufacture laser diodes for commercial sale. Low- to moderate-power lasers are made by Laser Diode Laboratories, RCA, and Texas Instruments.* Monsanto no longer manufactures lasers, but according to its latest data sheets still stocks several types. Sperry makes a relatively low cost laser, but production was halted as this article was being written. High-power single laser diodes are made by IBM and RCA. High-power arrays which generally do not require very high driving currents are made by Laser Diode Laboratories and RCA.

For practical communications experiments a simple lens arrangement is neces-

*Injection lasers are manufactured by: IBM Corporation, Federal Systems Division, 18100 Frederick Pike, Gaithersburg, Maryland 20760; Laser Diode Laboratories, 205 Forrest Street, Metuchen, New Jersey 08840; Monsanto Company, Electronic Special Products Division, 10131 Bubb Road, Cupertino, California 95014; RCA Electronic Components, 415 South Fifth Street, Harrison, New Jersey 07029; and Texas Instruments, Inc., Semiconductor Components Division, Post Office Box 5474, Dallas, Texas 75222.

sary to collimate the laser beam. RCA has an excellent brochure[6] which touches on optics. Edmund Scientific Company (600 Edscorp Building, Barrington, New Jersey 08007) stocks a wide variety of plano-convex and double-convex lenses which will permit the fairly broad beam from a laser to be narrowed to a few tenths of a degree. To capture all the light emitted by a typical laser choose a lens whose focal length is approximately equal to its diameter (near f/1 as possible).

Laser receivers are discussed in some detail by Campbell in his earlier article on injection lasers. There are numerous technical articles and references on the subject of detecting light – but keep in mind that the laser's light pulse peaks in the infrared and is quite narrow in time. Silicon detectors are admirably matched to GaAs lasers spectrally, and special types respond well to very fast pulse widths.

conclusion

Very few radio amateurs are experimenting with lasers of any kind, much less semiconductor lasers. The field is young – essentially wide open – and the time is ripe for amateurs with a pioneering instinct to become involved in the field of optical communications. The challenges are many (working with invisible light takes getting used to) but the rewards are great. In fact, it is quite likely that radio amateurs will make important contributions to this very important field of communications.

references

1. R. Campbell, W4KAE, "Injection Laser Experiments," *ham radio,* November, 1970, page 28.
2. R. Hunt, "Laser Diodes Need High Current Drivers," *Electronic Design,* September 27, 1970, page 50.
3. H. Brown, *et al,* "Avalanche Transistors Drive Laser Diodes Hard and Fast," *Electronics,* November 14, 1966, page 137.
4. N. Sullivan, "2000-A Pulse Generator," *Review of Scientific Instruments,* May, 1964.
5. W. Koechner, "Extremely Small and Simple Pulse Generator for Injection Lasers," *Review of Scientific Instruments,* January, 1967.
6. RCA, "Gallium Arsenide Lasers and Emitters," RCA Pamphlet OPT-100.

ham radio

mobile fm sequential encoder

This sequential encoder provides complete dial control of your local fm repeater

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A stable pulsed encoder for controlling fm repeaters is a relatively simple device requiring only a few components and a telephone dial. Since more and more fm repeaters are using pulse-type decoders and additional control it is convenient to have a small pulse encoder on hand whenever you need it.

The pulsed encoder in **fig. 1** uses lowcost parts, is easy to build, and provides an excellent sequential encoder for mobile fm operation. The full-size printed-circuit board in **fig. 2** reduces construction time to a few hours.

All components except the toroid oscillator coil should be available at your local parts store. Although I used a 2-Henry toroid in the circuit in **fig. 1** the surplus 88-mH toroids used by RTTY enthusiasts will do an excellent job if you juggle the values of the tone determining capacitors.

The circuit in **fig. 1** consists of a 2805-Hz oscillator (Q1), buffer stage (Q2) and emitter follower (Q3). The oscillator tone is set to the desired frequency by adjusting the .068 μF capacitor marked with an asterisk. In some cases it may be necessary to put two capacitors in parallel to obtain the desired tone.

The 3-second hold circuit (Q4) provides a 3-second carrier and tone at the end of the pulsing sequence. This allows for selective calling of a control pulse with "hands off" capability. I used an Allied 12-Vdc dpdt relay with a coil resistance of approximately 560 ohms at K1. This is not critical however, and almost any 12-Vdc relay will work. Transistor Q4 heats up a little, but since

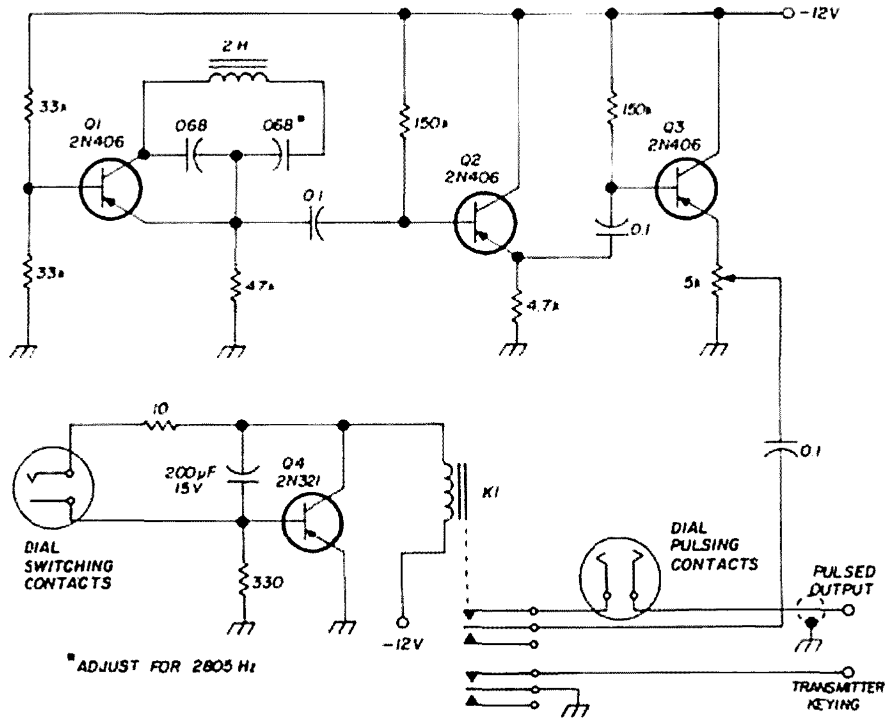


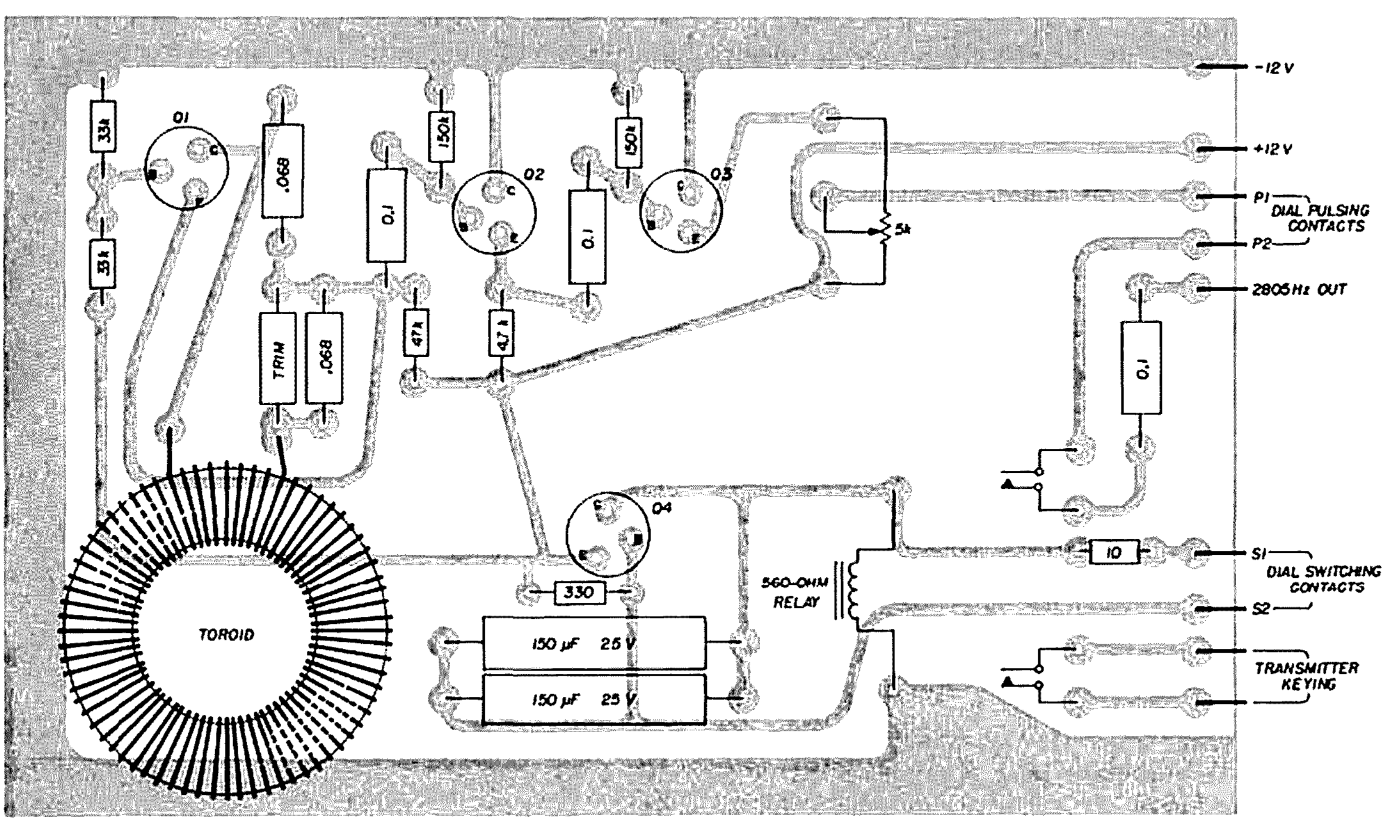
fig. 1. Circuit for the portable sequential encoder. Relay K1 is a 12-Vdc relay with dc resistance between 240 and 560 ohms.

control is not continuous it is not necessary to use a heavy-duty transistor.

If you use this circuit without a separate power supply (a 9-volt transistor radio battery will work), you must bypass circuit ground or *tone low* through a 0.1 μF capacitor to isolate the positive voltage from system ground. Since the circuit is open (no tone) until it is dialed and the relay picks up, the unit can be connected directly across the microphone input of your transmitter. The output matches the inputs of most fm transceivers (except those with carbon microphones).

ham radio

fig. 2. Full-size printed-circuit board and component layout for the sequential fm encoder.



a stable, variable-output weak-signal source

Receiver alignment generator, frequency standard, neutralization signal source are just a few uses for this instrument

Bruce Clark, K6JYO, 1019 El Dorado Drive, Fullerton, California 92632

A **necessary adjunct** to any test-equipment bench is a stable weak-signal source. This device is basically a crystal-controlled signal generator with an adjustable output. Its applications are analogous to the tunable signal generator, but it can be a much more useful tool than either a signal generator or a crystal calibrator by combining the properties of both. Several vhf signal sources have been described in *ham radio*[1,2] and *VHFer.*[3] All either lack frequency stability as the output is varied, or else they have too low output at the higher harmonic frequencies. The unit shown in **fig. 1** uses a pair of inexpensive RCA 40673 dual-gate mosfets to eliminate these shortcomings.

circuit description

The addition of a crystal-selector switch, S1, allows the choice of frequencies from 1 through 10 MHz. Presently, my unit uses 1 and 3.5 MHz for the high-frequency range and 8.0 MHz, 8.222 MHz, and 8.333 MHz for frequencies from 50 through 1296 MHz. High harmonic output is assured by the use of an H-P 2800 hot-carrier diode multiplier. Usable harmonic energy up to 1296 MHz been observed on a spectrum analyzer. With the circuit shown, the output was -82 dBm at 1296 MHz, a readily detectable level. A plot of output vs. frequency is shown in **table 1.**

Basically, the circuit is an untuned Colpitts crystal oscillator driving a wideband buffer, which in turn drives the hot-carrier diode multiplier. Two pots are used to adjust the output level. R2 sets the voltage level of G2 in the oscillator. This control is set to either of two calibrated points depending on the application. For example, if weak signal testing is desired, R2 is adjusted to the point at which oscillation is barely sustained. If a strong signal is desired, R2 is set to maximum output. The output-adjust pot, R7, sets the voltage of G2 on the 40673 buffer. The pot shown gives a -40 dB change in output level. This adjustment range can be calibrated in conjunction with a known attenuator and your receiver S-meter. Another approach is to mark off a 10-dB range in 1 dB steps on R7's scale and use an external attenuator for coarse adjustments.

construction

I used double-sided copper circuit board soldered together to form a 3 x 6 x 1 1/2 inch box. An LMB or Bud aluminum box of equivalent size could be used and the circuit built on a copper

sub-chassis inside as shown in fig. 2. However, rf leakage may be a problem at low levels using this construction. Extra screws must be used to keep the box rf tight. This will minimize leakage for most applications.

A simple ac supply capable of supplying 12-15 V at 25 mA may be built in the box; or if desired, a 12-15 V battery supply may be substituted. If an ac supply is used, the line cord should be well bypassed with feedthrough capacitors to eliminate rf leakage from this source.

applications

The uses of the unit are many. With a 1- or 3.5-MHz crystal, it makes an excellent frequency standard, band-edge marker, or receiver alignment generator. With an 8-MHz crystal, it can serve as a weak-signal source for vhf front-end alignment. In its maximum power output mode, the unit becomes a signal source for neutralizing converters and high-power rf amplifiers.

Other applications include calibration of receiver S-meters and determination of receiver dynamic (agc) range. Also, you could probably replace the harmonic generator with a tuned circuit, key the buffer, and go super QRP!

vhf converter alignment

A word of caution about the use of the signal source for vhf front end align-

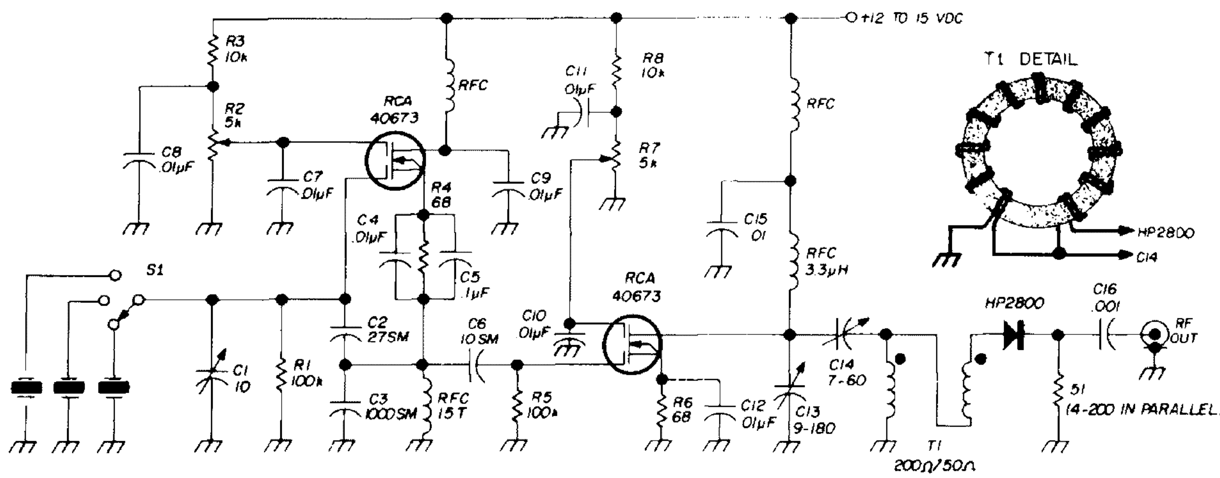


fig. 1. Schematic of the weak-signal source. T1: 7 turns no. 28 or 30 wire, bifilar-wound on no. 3E2A Ferroxcube toroid form. Rf chokes: 10 turns no. 24 or 26 on same form as T1 unless specified otherwise. C13: Arco 423. C14: Arco 404.

table 1. Output vs frequency.

frequency (MHz)	harmonic	output level (dBm)
8	-	-2*
16	2	-3.4
32	4	-17.4
48	6	-24.4
64	8	-29.0
144	18	-44.0
216	27	-51.2
432	54	-63.2
1296	162	-82.3

*0.625 mW across 50 ohms

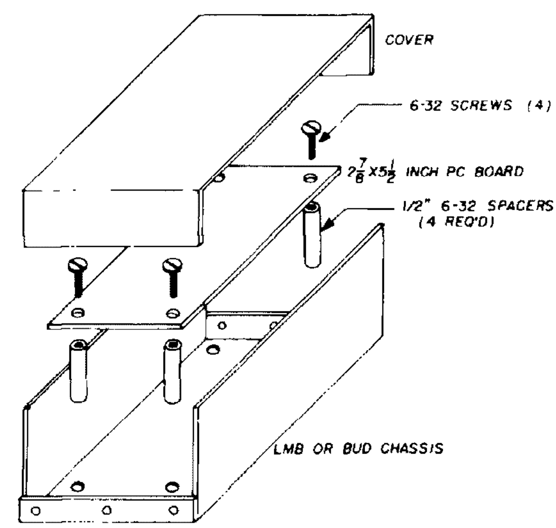


fig. 2. Alternate chassis design. An rf-tight enclosure is a must (see text).

ment: *if the* device under test (**fig. 3**) has little front-end selectivity or is badly mistuned, it may be possible to tune it up on the wrong harmonic of the signal source; e. g., a 432-MHz converter with a 28-MHz i-f should be tuned up on its 54th harmonic (assuming an 8.000-MHz crystal in the source). However, the 47th harmonic of 8.00 MHz is 376 MHz, which is the image frequency of the converter. *Since this is a lower-order harmonic, it* contains more energy than the desired one and could be the source of a lot of frustration. A similar situation exists with a 144-MHz converter using a 28-MHz i-f where the 18th harmonic is 144 MHz and the 11th harmonic is 88 MHz, which is the image. The solution in these cases is to include some selectivity between the signal source and the unit under test, at least until you're sure the converter is properly adjusted to the desired frequency. A second method is to offset the crystal frequency a few hundred Hz high using C1. The lower-order harmonic on the image frequency will then fall below the desired harmonic and can be easily identified.

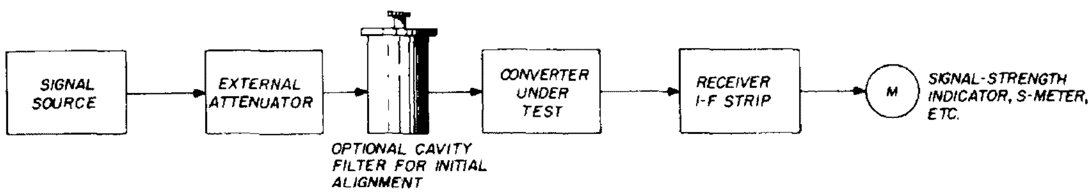


fig. 3. Setup for alignment of vhf converters.

neutralization of rf amplifiers

Fig. 4 shows the arrangement for neutralizing a receiver or transmitter rf amplifier. The procedure for receiver amplifier neutralization is as follows:

1. With B+ off disable the amplifier by opening the source bias resistor or opening a filament lead.

2. With signal source at maximum output, tune in signal on receiver.

3. Peak input and output circuits, then adjust neutralizing control for a null. Repeat several times.

4. Reconnect source bias resistor of filament lead.

For transmitter power amplifiers:

1. For tube rigs, *turn off B+ and screen voltage.* For transistor transmitters, leave B+ on and open emitter lead.

2. Connect weak-signal source to input. Adjust to maximum signal level.

3. Connect receiver to rf output of amplifier.

4. Tune in signal source and peak amplifier input and output controls for maximum S-meter indication on receiver.

5. Adjust neutralizing capacitor for minimum signal feedthrough. Repeat several times for maximum null.

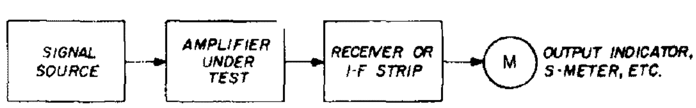


fig. 4. Arrangement for neutralizing receiver or transmitter rf amplifiers. Procedures are given in the text.

references

1. Del Crowell, K6RIL, "A 432 and 1296 MHz Signal Source," *ham radio,* September, 1969, p. 20.
2. J. W. Brannin, K6JC, "A Stable Small-Signal Source for 432 MHz," *ham radio,* March, 1970, p. 58.
3. S. Freeley, K6HMS, "A 144 MHz Weak-Signal Source," *VHF'er,* July, 1965.

ham radio

multimode i-f system

External refinements for older receivers — featuring ssb, fm, and synchronous a-m detection plus a wide-range audio compressor

Richard Factor, WA2IKL, 30 West 60th Street, New York, New York 10023

Here's an outboard i-f and audio system that can be used with older receivers to bring them up to today's standards. Most older sets have little room to spare for adding a selective filter and product detector — essential requirements for good ssb performance. In addition to *normal* ssb operation, the *circuit* described here provides:

1. An fm detector essentially immune to a-m interference.

2. An audio compressor with a wide dynamic range and slow time constant, which eliminates constant riding of the volume control during net operation and fm-channel monitoring.

3. Synchronous a-m detection.

the circuit

The multimode i-f and audio system schematic is shown in fig. **1.** Semiconductor enthusiasts will note an assortment of rather unusual ICs in the i-f and detector circuits. One of the new Signetics phase-locked loops, the PLL 560, is used as an f-m detector and carrier generator. A Motorola MC1596 balanced-modulator IC operates as a balanced demodulator in the product detector. The audio compressor, fig. **2,** uses a type 709 op amp with a Raysistor CK1112 as the gain-control element.

To understand how the i-f/detector works, refer to **fig. 3,** paying particular

attention to the switch positions for each mode.

single sideband reception

Sideband operation is similar to that in standard receivers: the bfo generates a signal equal to the (suppressed) carrier frequency of the signal you are tuning. The mechanical filter passband is above the bfo for usb; below it for lsb. The bfo output goes to the phase-locked-loop input, and the loop vco output goes to one input of the product detector. In this operation, the phase-locked loop acts as a buffer since the input frequency is equal to the output frequency, and the vco output is the only signal used. The mechanical filter output goes to the other product-detector input, and the output of the product detector is the reconstructed audio signal.

a-m reception

If you try to receive a-m with a sideband receiver, inevitably you run into the problem of precision tuning. In ssb reception a 50 Hz error isn't serious, but in a-m reception any error is a disaster. You have a carrier and a bfo signal. If a phase difference exists between them, their outputs may cancel, leaving no

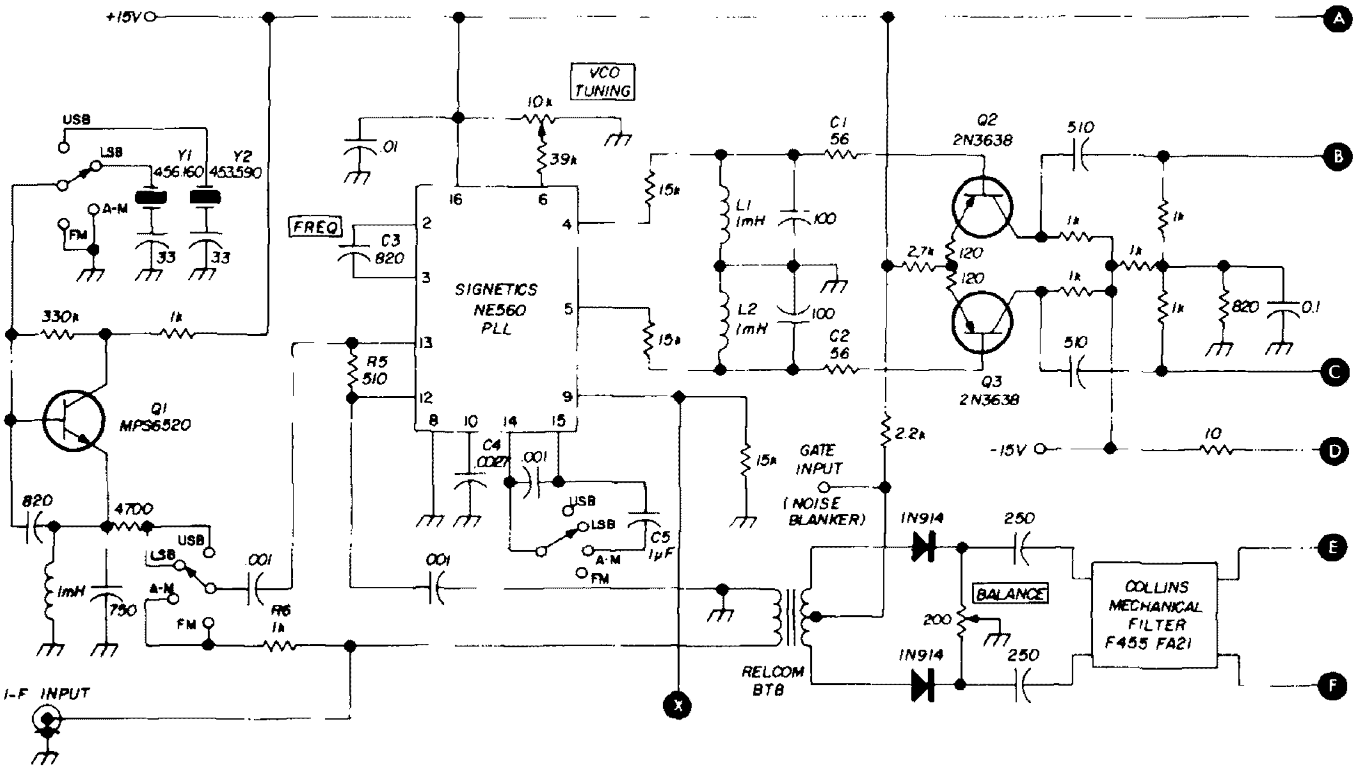


fig. 1. Schematic of the multimode i-f/audio system. Substitute devices may be used; e.g., the less-expensive PL 565 can replace the PLL 560. Also, equivalent IC's by Fairchild and Signetics are available for the balanced demodulator.

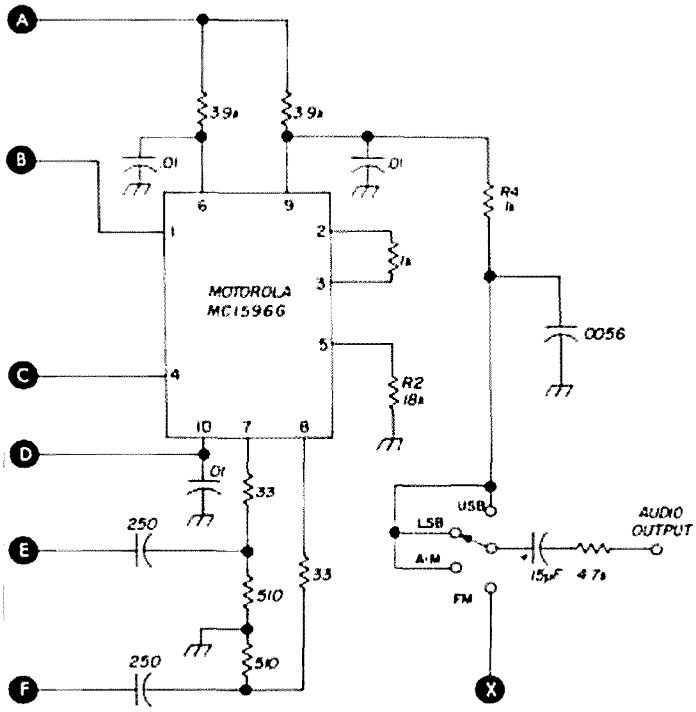
reference for demodulation. If there is a frequency error a flutter will occur as the phases change, or a beat note will occur if the phase difference is sufficient. Synchronous detection eliminates these problems.

The a-m signal is simultaneously applied to the mechanical filter input and the phase-locked-loop input. The phase-locked-loop synchronizes with the carrier of the a-m signal and provides a reference of the correct phase to demodulate the signal. Meanwhile, the signal is processed by the mechanical filter, which can be positioned anywhere with respect to the carrier frequency, thus improving selectivity.

f-m reception

The phase-locked loop synchronizes with the carrier of the incoming signal. If

the signal increases in frequency, a dc voltage produced by the loop increases the oscillator frequency. If the signal decreases in frequency, the dc voltage docreases to re-sync the oscillator. Thus,



the dc voltage is proportional to the input frequency. Producing a voltage proportional to the frequency of the input signal is called f-m detection. The allowable deviation of the f-m signal is dependent upon the receiver i-f bandwidth before the mechanical filter. This will usually be sufficient to copy downconverted two-meter signals.

compressor

The compressor consists of an operational amplifier with a variable resistor in the feedback loop. A low resistance decreases the gain. Since the "resistor" is light dependent, it can be controlled electronically. The op-amp output is rectified and applied to a long-time-constant RC network at the input to a Darlington stage. The Darlington output goes to the lamp half of the Raysistor. Thus, as the amplifier output tends to increase, the circuit gain decreases, maintaining a constant average output through a wide dynamic range.

circuit details

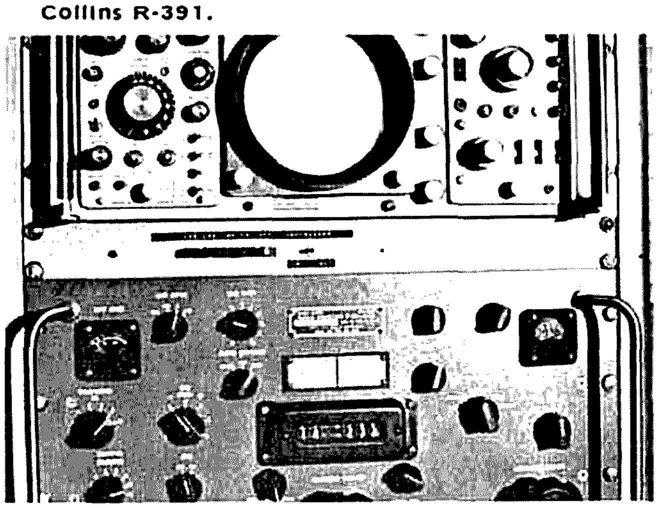
The i-f input from my receiver was about 1V p-p into 50 ohms. This was ideal for driving a 2-to-1 broadband transformer I had available. The transformer output is balanced, which makes it ideal for gating if you wish to use a i-f noise blanker.

Preliminary experiments indicated that a noise blanker inserted here offered little improvement over conventional limiters, probably due to the already narrow selectivity. So if you don't have such a transformer don't worry about it. Just connect the filter directly to the i-f output. The filter output is resonated by the two 250-pF capacitors and is connected differentially to the product-detector input. The two 33-ohm resistors are parasitic suppressors. R1 and R2 set the product detector gain and bias. R1 may be adjusted, depending upon the receiver i-f output. R2 should be adjusted if you wish to use a different power-supply voltage. R3 and R4 and the associated capacitors form a low-pass filter on the product-detector output to keep 455 kHz signals out of the rest of the circuit.

buffer and bias network adjustment

The assembly between the vco output and the product detector carrier input

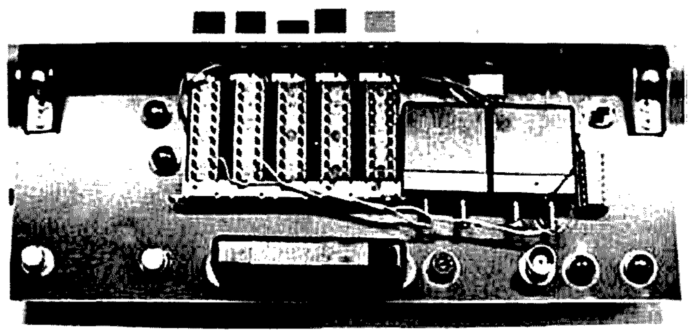
Front view of the external multimode i-f/audio and compressor system. Note the neat arrangement of the rack-and-panel unitized packaging. Receiver below the multimode system is a Collins R-391.



(containing L1, L2; Q2, Q3) is a rather elaborate buffer amplifier and biasing network. The vco output is a square wave with a positive dc bias. The product detector requires a sine wave with a negative dc bias. Differential amplifier Q2, Q3 make up for some of the signal lost in the resonant filters. By changing the resonant-filter capacitors C1 and C2, variable phase shift may be introduced if necessary. To align the phase, switch to a-m and insert a 455 kHz signal into the input. Compare the signals on pins 1 and 7 of the product detector. If they are in phase, or nearly so, no adjustment is necessary (likewise if they are 180 degrees out of phase). If they are approaching 90 (or 270) degrees, remove 30 pF from C1 and C2. This sounds a lot more critical than it really is. Even if the signals are *exactly* 90 degrees out of phase things rapidly improve with any introduction of phase shift.

phase-locked loop

The phase-locked loop has been adequately described[1] so I won't go into detail here. C3 sets the vco free-running frequency to about 455 kHz. The 10k tuning pot is a vernier control to center the frequency. R5 and R6 form an input divider network to avoid overloading the phase-locked loop and to increase a-m rejection in the f-m mode. C4 is a de-emphasis capacitor to lower high-frequency audio response of the phase-locked loop. C6 is a loop filter capacitor, which ensures stability of the loop and yet allows it to change frequency fast enough to follow the f-m input. In the a-m mode C5 is switched in to maintain a



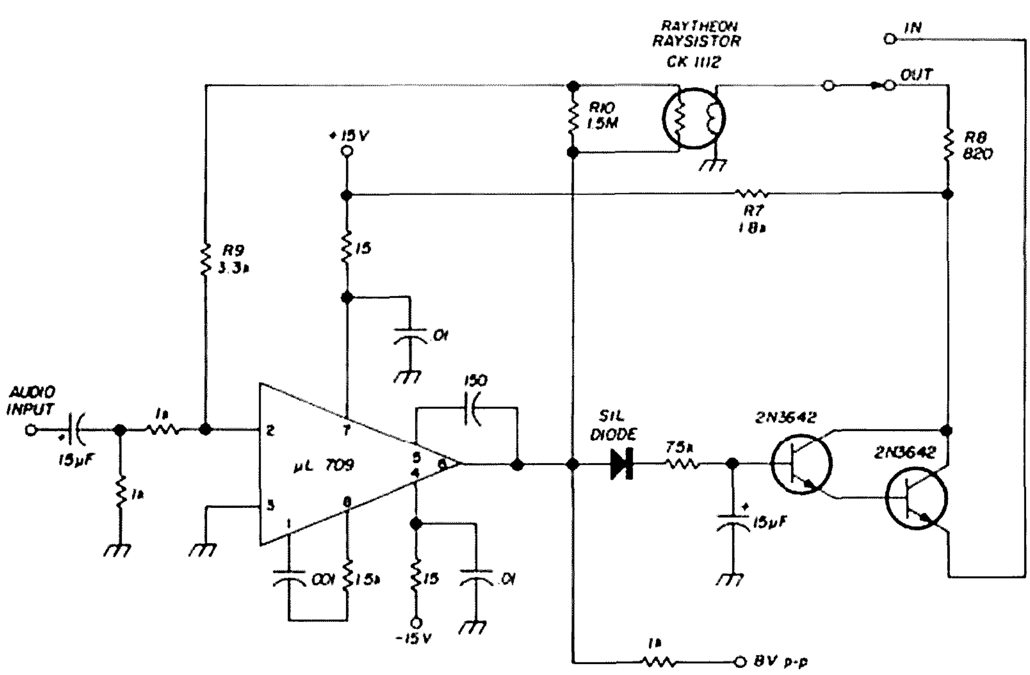
Parts layout and wiring of the final unit. Short leads are recommended for minimum intrastage rf coupling.

stable output frequency during modulation.

bfo

The bfo is a standard oscillator circuit. I never could remember who it was

fig. 2. Audio compressor schematic. Circuit is designed around the Raytheon CK1112 Raysistor, but a cadmium sulfide photo cell and low-current pilot light could be used as a sensor. See text.



named after. However, the important thing is that, if you have spent as many hours as I have trying to get those miserable FT243 crystals to oscillate, you'll be happy to have a decent circuit regardless of whom it is named after. One important note: keep the leads to the crystals *short* or this circuit won't work either.

construction

I built the final unit as shown in the photos. The bfo and input circuits are on one side of the filter; the product detector and audio circuits are on the other. Before building the final unit, I breadboarded the circuits. All the components had long leads, but both circuits seemed to perform equally well. However, leads should be short to keep rf radiation to a minimum. I used many bypass capacitors and parasitic suppressors. If you want to leave them out, do so at your peril. The inherently balanced circuitry probably contributed as much to stable and reliable operation as did the wiring technique.

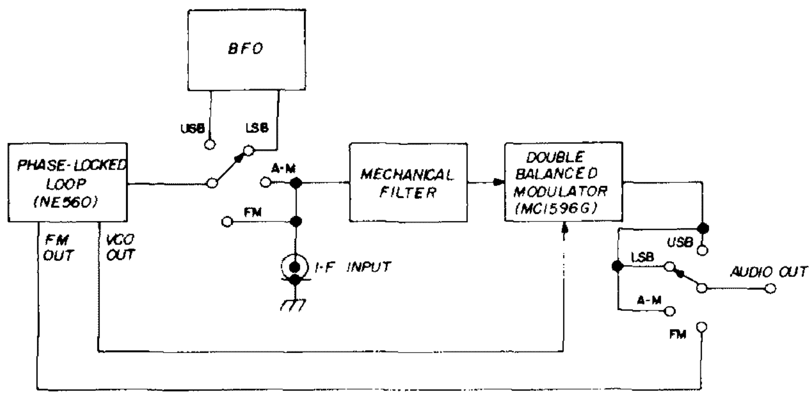


fig. 3. System block diagram.

compressor design

The Raytheon CK1112 Raysistor may be difficult to obtain. A Raysistor can be built for about a dollar, using a cadmium sulfide photocell and a low-current pilot light. R7 is determined by connecting it in series with the lamp and the power supply. Enough current should flow in the lamp to give a photocell resistance of 1 – 2k. Measure the lamp voltage under normal operation. R8 should be connected in series with R7 to the lamp when you don't want compression in this way, compression can be switched in and out without changing the circuit gain and without introducing transients in the audio circuit. R9 sets the minimum gain of the op amp with the photocell fully conducting. R10 sets the gain with the photocell off. The ratio of these resistors is approximately equal to the amount of compression available. The values I have used give a range of about 55 dB, which is somewhat extravagant since an audio signal-to-noise ratio of 55 dB is seldom encountered in ham work. Vary these resistors to suit your taste. They are relatively independent of the type of photocell you use.

parts availability

The PLL 560, which at this writing is fairly expensive, can be replaced by a PLL 565, which isn't. The balanced demodulator is manufactured by Motorola, Fairchild, and Signetics. The Collins mechanical filter can probably be replaced with a Japanese unit that I've seen advertised for much less money. The BT8 transformer and CK1112 Raysistor have already been discussed. Everything else is quite inexpensive. The money you save on parts might well pay for one of those old surplus receivers, which can be resurrected with the multimode i-f system described here.

reference

1. Jim Kyle, K5JKX, "The Phase-Locked Loop Comes of Age," *73,* October, 1970, page 42.

ham radio

simple transconductance tester for field-effect transistors

A versatile, low-cost test circuit for checking the performance of field-effect transistors

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At the last West Coast VHF Conference my favorite two-meter fet converter's noise figure was checked and found wanting. Since this converter[1] had previously been measured at better than 2 dB it was apparent that it needed some work. The signal generator, sweep generator and noise generator were broken out in an effort to restore the beast to its original operating condition. After many fruitless hours of alignment and checking the converter was only slightly improved and was still far below what it should have been.

Although I had replaced the fet in the first r-f stage I had not checked the other fets, but assumed that they were good since the converter was working. By accident (or for reasons I don't remember) I removed the second r-f amplifier fet from its socket. There was virtually no difference in received signals with or without the fet! The transistor was dead. With a new fet in the socket it took all of 15 minutes to realign the converter, restore the bandwidth to 4 MHz, and verify that the noise figure was once again better than 2dB.

A little analysis showed what was happening. The defective stage was a neutralized common-source amplifier. With a dead transistor the signal at the gate lead was coupled across the neutralizing inductance to the drain tank circuit and the mixer, less 10 dB of amplification.

Everyone knows the first step in

troubleshooting is to check the suspect tube or transistor. However, this requires a suitable tester, which I did not have. I have one now, though, that uses an external audio oscillator and ac voltmeter or oscilloscope (see **fig. 1**). Cost, even with all new components, is minimal.

calibrated oscilloscope connected across the output terminals indicates the output voltage developed across the 1k drain load, R3. The forward transconductance of the transistor under test, in micromhos, is equal to 10,000 times the output voltage. Thus, if the voltmeter

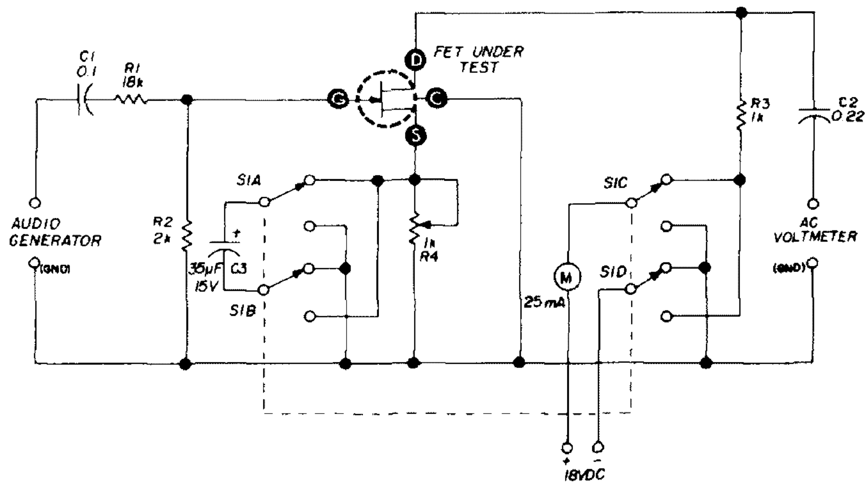


fig. 1. Circuit of the fet transconductance tester. Switch S1 is a 4-pole, double-throw, non-shorting rotary.

circuit

A 1-volt, 1000-Hz audio signal is applied to the input of the fet tester. R1 and R2 form a 10:1 voltage divider that reduces the signal at the gate of the fet under test to 100 millivolts; virtually no gate current will flow under zero-bias conditions. An external ac voltmeter or across the output terminals reads 0.4 volt, the transconductance of the device under test is 4000 micromhos.*

Potentiometer R4 is used to set the source bias and hence, the drain current of the fet. There is a wide variation in drain currents of individual transistors of the same type. The manufacturer usually specifies a normal operating drain current so it is helpful to monitor the drain current and set it to the desired value with R4. The milliammeter may be built into the tester or connected in series externally with the power supply.

Switch S1 reverses power-supply polarity so the power supply connections do not have to be reversed when changing from n- to p-channel devices. This switch also reverses the polarized source-bypass capacitor, C3. This capacitor is needed to prevent degeneration across R4 which would reduce the transconductance reading. If S1 is set to the wrong position no damage will occur to the transistor; the output reading will be zero, or very low. Reversing the switch position will result

*The explanation of this relationship is as follows. The forward transconductance, gfs, of the device under test may be expressed in mhos as

$$g_{fs} = \frac{i_d}{e_g}$$

where i_d is the a-c component of the drain current (in amperes) and e_g is the a-c gate voltage. From Ohm's law,

$$i_d = \frac{e_o}{R_d}$$

where e_o is the a-c output voltage and R_d is the drain load resistance (designated R3 in fig. 1). By substitution,

$$g_{fs} = \frac{e_o}{e_g R_d}$$

If $e_g = 0.1$ volt and $R_d = 1000$ ohms, the conditions under which this circuit operates, then

$$g_{fs} = \frac{e_o}{100} \text{ (mhos) or } 10{,}000\, e_o \text{ (micromhos)}$$

in a substantially higher reading if the transistor is good. Thus, the tester will also indicate whether the fet is a p- or n-channel type. In addition to testing junction fets, the fet tester will also check *single-gate* depletion-type mosfets.

construction

The fet tester may be built in, or on, virtually anything. I used an old plastic box; this permitted me to use ordinary 6-32 screws as input, output and power supply terminals without additional insulation. Lead length and physical layout are not at all critical; the tester can truly be built from junk-box parts.

Since there is absolutely no standard basing configuration for the many types of fets on the market the test socket may be wired according to personal preference. However, be sure to designate on the tester which pin is which so that you can insert the fet properly.

external requirements

Although the voltage of the external power supply is not critical 18 volts has been specified for two reasons: 18 volts is high enough to keep the drain potential above the knee of the E_D-I_D curve if the device under test draws sufficient drain current to cause an appreciable drop across R3; 18 volts is below the normal breakdown voltages specified for most transistors.

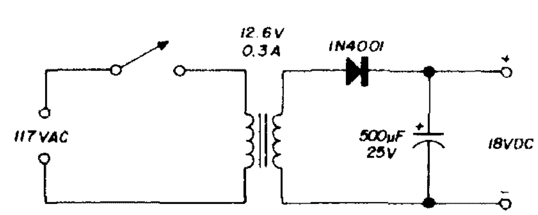


fig. 2. Suggested power supply for the fet tester.

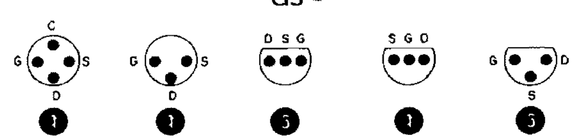
The circuit for a suitable ac power supply is shown in **fig. 2.** This could be incorporated into the tester if desired. However, it is probably easier and cheaper to use two 9-volt transistor-radio batteries. Since the current drain is intermittent and rarely exceeds 15 mA they should last their entire shelf life.

The audio oscillator must supply a reasonably good sine wave at 1 volt rms

table 1. Characteristics of popular field-effect transistors.

type	channel	base diagram	transconductance (µmho)*	I_{DSS} (mA
2N3823	N	1	3500-6500	4-20
2N4360	P	2	2000-8000	3-30
2N4416	N	1	4500-7500	5-15
2N5397	N	1	6000-10,000	10-30
2N5398	N	1	5500-10,000	5-40
E300	N	2	4500-9000	6-30
MPF102	N	3	2000-7500	2-20
MPF103	N	3	1000-5000	1-5
MPF104/2N5458	N	3	1500-5500	2-9
MPF105/2N5459	N	3	2000-6000	4-16
MPF106/2N5485	N	3	2500-6000	4-10
MPF107/2N5486	N	3	4000-8000	8-20
MPF108	N	3	2000-7500	0.5-24
MPF109	N	3	800-6000	0.5-24
TIS34/2N5248	N	4	3500-6500	4-20
TIS88/2N5245	N	5	4500-7500	5-15
TIXM12	P	2	5000-20,000	5-25
UC734	N	1	3500-6500	4-20

*Transconductance at $V_{GS} = 0$.



into a high-impedance load. The frequency of 1000 Hz was selected because many manufacturers specify forward transconductance at that frequency. Actually, any audio frequency will serve, provided the capacitors in the circuit are increased proportionately if a lower frequency is used. For example, if you decide to use a 60-Hz source, the values of C1, C2 and C3 must be increased roughly by the ratio of the frequencies, 1000/60, or approximately 16 times.

operation

Operating the fet tester is very simple. Connect the audio oscillator to the input terminals and set the audio voltage to exactly 1 volt. Connect the ac voltmeter (or oscilloscope) to the output terminals; if an external power supply is being used

connect it to the dc supply terminals. Set switch S1 to the position which corresponds to the fet to be tested – n or p channel.

Put the fet in the socket, being careful to match its leads to the socket pin designations. It should be noted that junction fets are symmetrical devices so drain and source leads can be interchanged with no difference in performance. The gate lead, however, must be in the proper socket hole.

To check a depletion-type mosfet be sure to keep all of its leads shorted together with a strand of wire before it is inserted or removed from the socket. Also, be certain that the bulk or substrate lead, which is generally common to the case, is properly connected.

Adjust R4 for the desired drain current and read the output voltage. Multiply by 10,000 and you have the forward transconductance in micromhos. If the output voltage is abnormally low, reverse the position of S1. If you still do not get a normal reading on the ac voltmeter, and the transistor is properly connected, the transistor is defective. Conversely, switch S1 may be used to determine whether the transistor is a p- or n-channel type. The switch position which results in the *lower* dc drain current indicates the type.

Table 1 lists some of the more common fets with their transconductance range, zero-bias drain current (I_{DSS}) and basing configuration. Note that the transconductance is specified at zero gate bias (V_{GS}) which results in I_{DSS}. The higher the drain current, the higher the transconductance.

Although the fet tester will probably be used infrequently, as is true for any transistor or tube checker, the small amount of time, effort and money required to build it can pay off handsomely (just reread the first three paragraphs of this article).

reference

1. William I. Orr, W6SAI, "Radio Handbook," 18th edition, Editors and Engineers, New Augusta, Indiana, pages 624-629.

ham radio

automatic frequency control for receiving RTTY

This circuit cures problems with drifting RTTY signals and is particularly useful with drift-prone receivers

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The **automatic frequency-control** circuit described here is the result of trying to copy the Weather Bureau's RTTY station on 14.395 MHz with a not-so-stable communications receiver. After trying to stabilize the receiver so my ST-5 demodulator[1] would copy unattended for a reasonable length of time it began to look as though an afc system might be easier. The circuit in **fig. 1** is simple and uses low-cost integrated-circuit operational amplifiers and surplus 44-mH toroids.

circuit operation

Since it's necessary to present a 2975 Hz tone to the 2975 tuned circuit in the demodulator, 2975 Hz looked like a good standard for frequency control. In this circuit U1 amplifies the 2975 Hz signal that is intermittently present during reception. A high input impedance prevents loading the demodulator tuned circuit.

The IC op amp is set for a gain of about 2.7; the output is coupled to a pair of tuned circuits. One of the circuits is tuned to about 150 Hz above 2975; the other is tuned about 150 Hz below. The rectified dc output voltages are opposite and cancel when the frequency is about midway (2975 Hz).

As the input freguency goes up the voltage goes negative; as the frequency goes down the circuit provides a more positive output. However, the voltage swing is not large enough for control purposes, so another op amp must be used as a dc amplifier. These amplifiers provide an output that swings from positive to negative with respect to ground, duplicating the polarity of the discriminator output. The output voltage from U2 is

fed to a varactor or tuning diode installed in the receiver's tuning oscillator (fig. 2).

When the frequency drifts so the RTTY tone goes up, the discriminator output goes more negative; tuning diode operating point properly; the bias control (R2) adjusts bias through the inverting input of U2. Because of the high resistance in the non-inverting input (pin 3) the necessary bias is slightly negative.

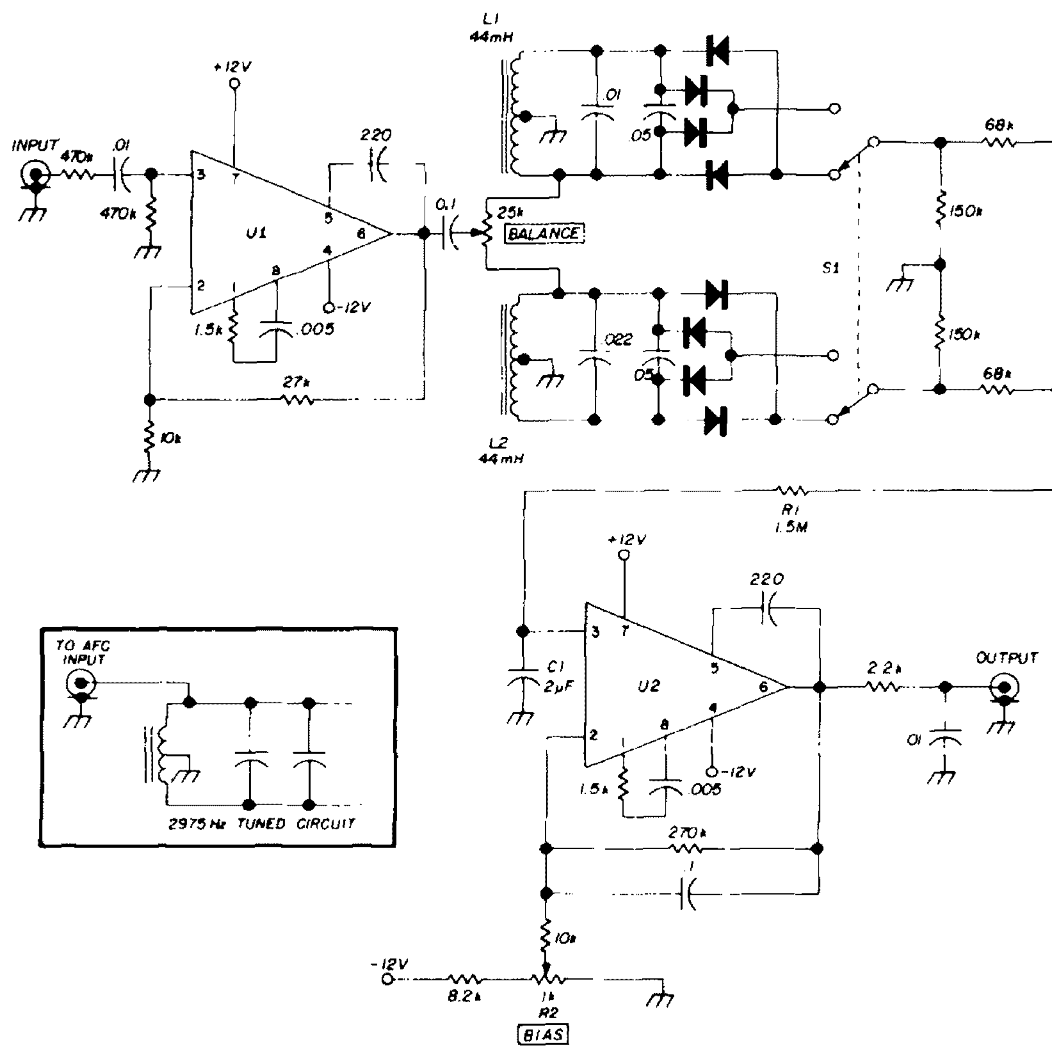


fig. 1. Circuit for the RTTY afc unit. Discriminator diodes are germanium types such as 1N34A. Integrated circuits U1 and U2 are 709 types: Motorla MC1709CG, TI SN72709L, Fairchild uL709C, etc. Inset shows 2975-Hz signal pickup in RTTY demodulator.

capacitance decreases, increasing the oscillator frequency to bring the audio tone back to 2975 Hz.

In most RTTY signals 2975 Hz is used as the mark tone. Therefore, it is present for a larger percentage of time than the space signal. Since the tone is constantly on and off a simple RC filter (R1, C1) presents a long time constant to the pulsating dc and smooths the output to prevent warble of the receiver audio.

The varactor must be biased to set its Two units were tried; one required −0.4 V, the other, −0.35 V.

initial setup

After a thorough wiring check connect the unit to the receiver and RTTY demodulator. (Power from the ST-5 plus and minus power supply may be used since total current drain is only 5 mA. If the unit is used with another demodulator use any well filtered power supply, or batteries.)

The afc circuit is connected to the communications receiver through capacitor C2 (fig. 2); connect it to the cathode of the oscillator tube (or emitter of a transistor oscillator). Connections to high-impedance points in the oscillator circuit may result in large changes in dial

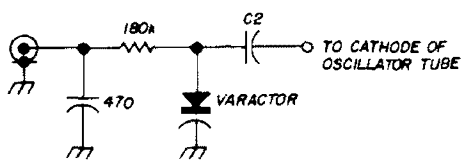


fig. 2. Varactor tuning circuit connects to the cathode of the oscillator tube. The varactor is not critical; it may be a silicon rectifier such as the 1N4383 or a small 20-pF voltage-variable diode. See text for selecting proper value for C2.

calibration. If this is necessary a very small value should be used at C2. Capacitance should be slightly less than that which causes a noticeable audio warble or chirp when a RTTY signal is received. A 5- or 10-pF capacitor is a good starting point.

adjustment

For proper adjustment, connect a vtvm across the output jack and make the following adjustments:

1. Disconnect the input cable from the demodulator.
2. Adjust the *bias* control to -5 volts output.
3. Connect the demodulator to an accurate 2975-Hz tone.
4. Adjust the *balance* control to bring the output back to -5 volts.
5. Remove the 2975-Hz tone and connect the demodulator to the receiver.

When a RTTY signal is properly tuned the vtvm will read -5 volts. This reading will go up or down as you tune across the signal. The 10-volt scale makes a fine tuning indicator with -5 volts as zero center. If you tune slowly the audio RTTY tones will remain nearly the same.

Depending on the relative positions of the bfo and the receiver oscillator it may be necessary to reverse S1. If the polarity is wrong for your receiver the afc unit will *prevent* you from tuning in a RTTY signal!

The voltage at either end of the 2975-Hz coil in the ST-5 demodulator is on the order of 12 volts peak-to-peak. When using another type of RTTY demodulator the input signal to U1 should be approximately at this level. This may be accomplished by lowering the resistance between pin 3, U1, and ground if the signal is too large. If the signal is too low the gain of the first op amp may be increased by increasing the value of the 27k feedback resistor between pins 6 and 2 of U1.

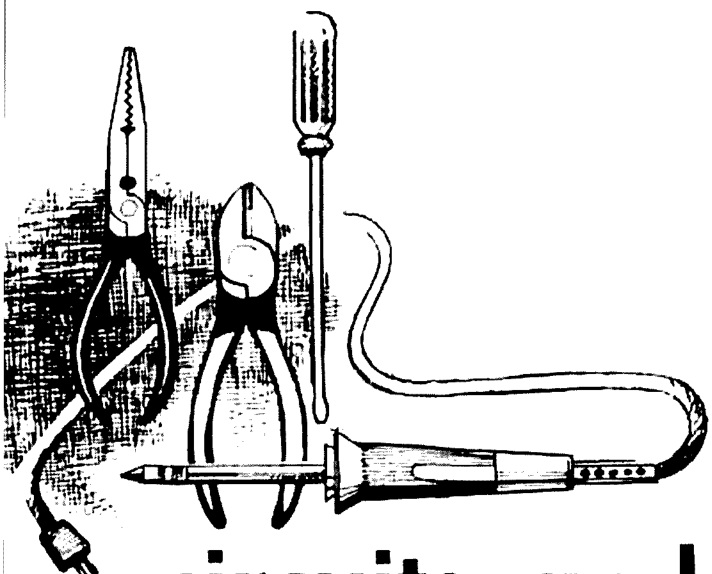
This afc unit has proven very useful in my station. It allows the use of an inexpensive receiver to monitor a RTTY station for long periods of time without retuning to compensate for receiver or transmitter drift.

reference

1. Irving M. Hoff, W6FFC, "The Mainline ST-5 RTTY Demodulator," *ham radio*, September, 1970, page 14.

ham radio

"Edward, you're fouling up the TV again!"



circuits and techniques

ed noll, W3FQJ

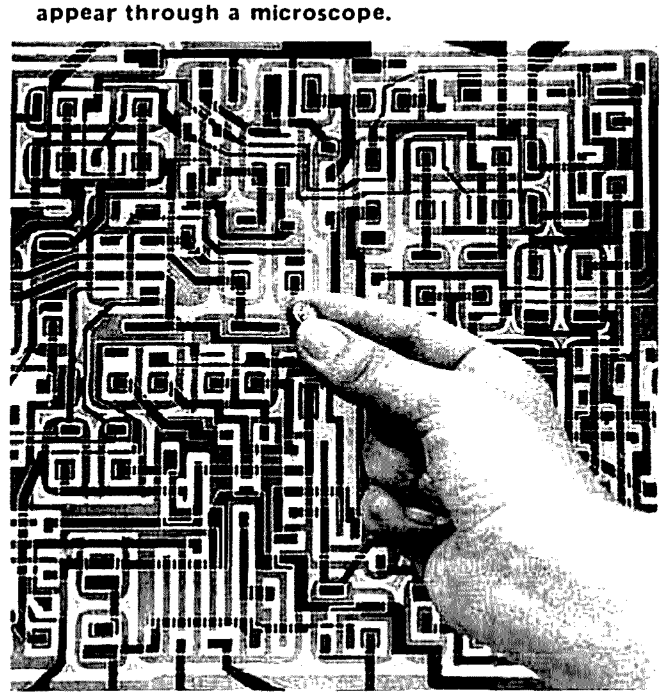
phase-locked loops

Do you wish that your local oscillator would hold precisely on frequency after you have tuned in an a-m, fm or fsk signal? How would you like to automatically track an incoming signal that drifts in frequency? How would you like to demodulate that signal even though it is plastered with noise? For an affirmative answer to any of these questions you must enter the realm of the phase-locked loop. A fine discussion of the basic theory of this circuit was presented recently by VE5FP.[1]

Although the phase-locked loop concept has been around since the early 1930s its high complexity has limited its use to specialized applications in expensive military and avionics equipment. For an idea of the complexity of this circuit consider an amateur radio version of the phase-locked loop described by W2CRR and WØAHN[2] that used a total of 16 vacuum-tube stages or VE5FP's solid-state phase-locked oscillator that uses 7 transistors plus a large quantity of resistors, capacitors and inductors. It's no wonder that this versatile circuit has seen little application in amateur radio equipment.

The integrated-circuit versions of the phase-lock loop are as complex as their discrete component counterparts. The Signetics NE560 for example, the first phase-locked loop IC, contains no less than 25 transistors, 7 diodes and 27 resistors. With this device it is possible to build high performance fm detectors, RTTY demodulators, signal-tracking filters, stable local oscillators, frequency synthesizers and frequency multipliers as well as more complex systems. Additional components included in its sister unit, the Signetics NE561B, provide capability for operation as a frequency-selective (coherent) a-m detector. These monolithic phase-locked loops have a frequency range from 1 hertz to 15 MHz without the need for tuned circuits.

Signetics integrated-circuit IC is dot at the center of the package held here in the fingers. Background is greatly magnified photgraph of the circuit as it would appear through a microscope.



applications

There are several applications for phase-locked loop ICs in amateur communications receivers. They can be used as small efficient synchronous a-m detectors with excellent linearity and noise immunity. Can they make vhf-uhf a-m DXing a more realistic endeavor? How about tuning in and holding a-m signals on bands crowded with other signals? I'll soon be checking one out on 160 meters.

For fm the phase-locked loop IC may be used in compact i-f, limiter and demodulation systems. The IC's wide frequency range and no-tuning characteristic are especially attractive in receiver designs that use more than one i-f frequency.

How about a direct-conversion fm receiver? For 10-meter operation this is probably possible today using off-the-shelf phase-locked loop ICs. If you already have a high-quality communications receiver to which you would like to add an fm demodulator, the phase-locked loop IC may be your answer; it is an efficient demodulator that will accomodate any i-f frequency up to 15 MHz. (A multimode i-f system for fm, a-m and ssb which is designed around the Signetics NE560 is described by WA2IKL on page 39 of this issue).

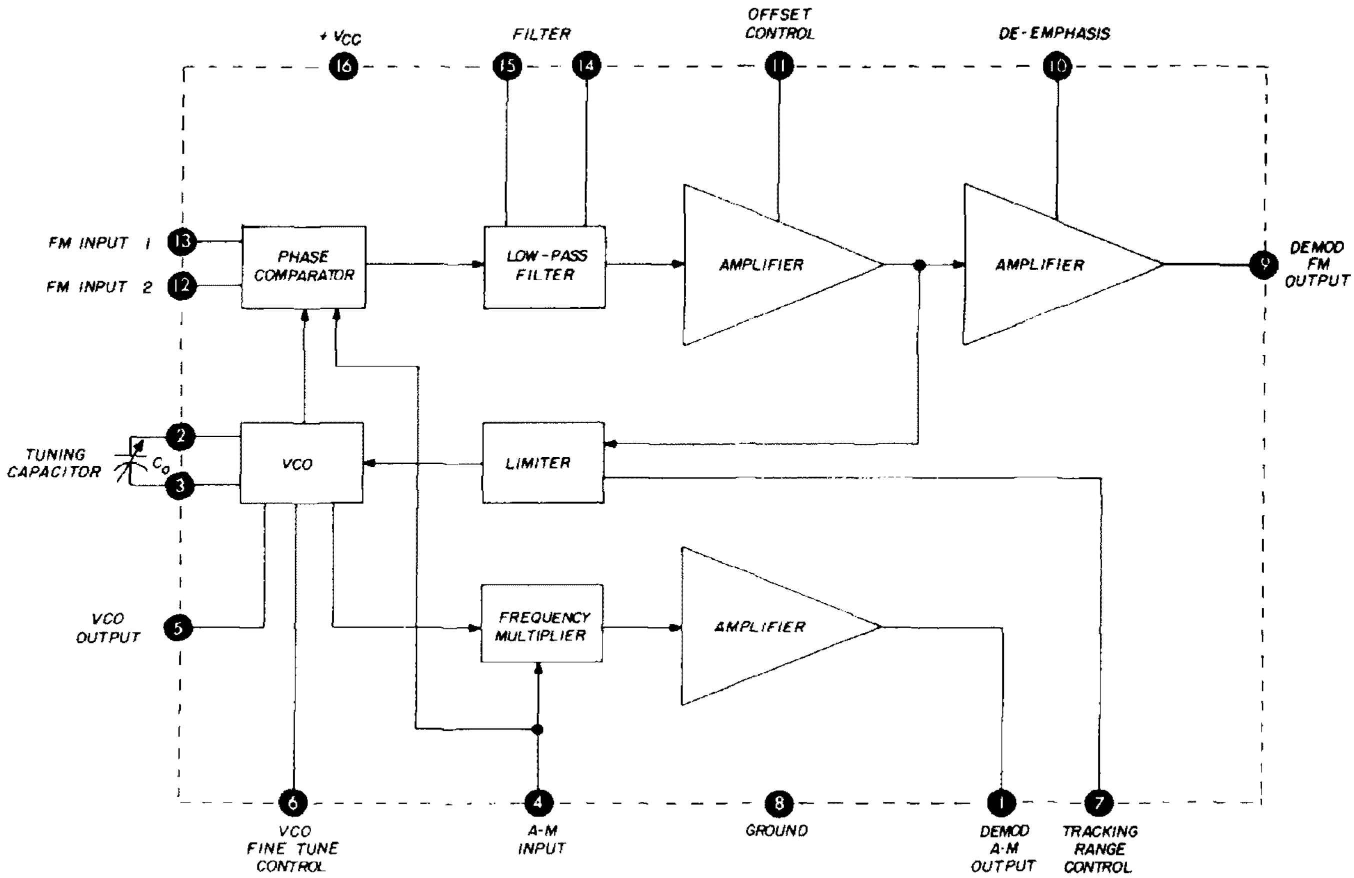


fig. 1. Block diagram of the Signetics NE561B phase-locked loop integrated circuit.

Application of the phase-lock loop concept will also provide a stable vfo or frequency synthesizer for amateur fm operation. The phase-locked loop can also be used to advantage in an a-m receiver that will hold precisely on a net frequency and still follow those stations that just can't seem to find the right spot.

In addition, the phase-locked loop IC operates well as a frequency multiplier or a frequency divider. With a stable crystal reference oscillator or vfo, the vco in the IC can be operated on a harmonic or sub-harmonic of the reference frequency. These capabilities encourage the design of

versatile frequency calibrators and high-stability multiband vfos.

phase-locked loop ic

A block diagram of the Signetics NE561B phase-locked loop IC is shown in fig. 1. An incoming fm signal is applied to the phase comparator through pins 12 and 13. A reference signal from the voltage-controlled oscillator is applied internally. The vco frequency is adjusted externally with a variable capacitor between pins 2 and 3, or with a dc control voltage connected to pin 6. The vco output is available at pin 5.

The output of the phase comparator is applied to a low-pass filter. The filter characteristics are controlled with an external network connected between pins 14 and 15. A two-stage amplifier follows the low-pass filter. The de-emphasis characteristic is set with a capacitor from pin 10 to ground. The demodulated fm output is available at pin 9.

The dc error voltage at the output of the first amplifier is applied to the limiter. This filtered error voltage is applied to the vco, causing it to phase lock to the incoming signal. When the incoming fm signal deviates in frequency with modulation the error voltage developed at the output of the phase comparator corresponds to the demodulated output.

When a phase-locked loop locks on an fsk signal the error voltage from the phase comparator is in discrete voltage steps that correspond to mark and space, or the digital 1 and 0.

A-m demodulation (coherent detection) circuit operation uses all the fm demodulation circuits *plus* a multiplier which functions as a mixing-type a-m demodulator. The incoming a-m signal is applied to the mixer input at pin 4. The IC locks on the a-m carrier frequency; in fact, the output of the vco is the same frequency as the carrier but without modulation.

The vco signal is also applied to the multiplier. A low-pass filter at the output of the multiplier cancels the rf carrier and sideband components and produces a difference signal corresponding to the a-m modulation. This system of demodulation is referred to as phase-lock a-m detection. It is much less subject to noise than conventional a-m detectors.

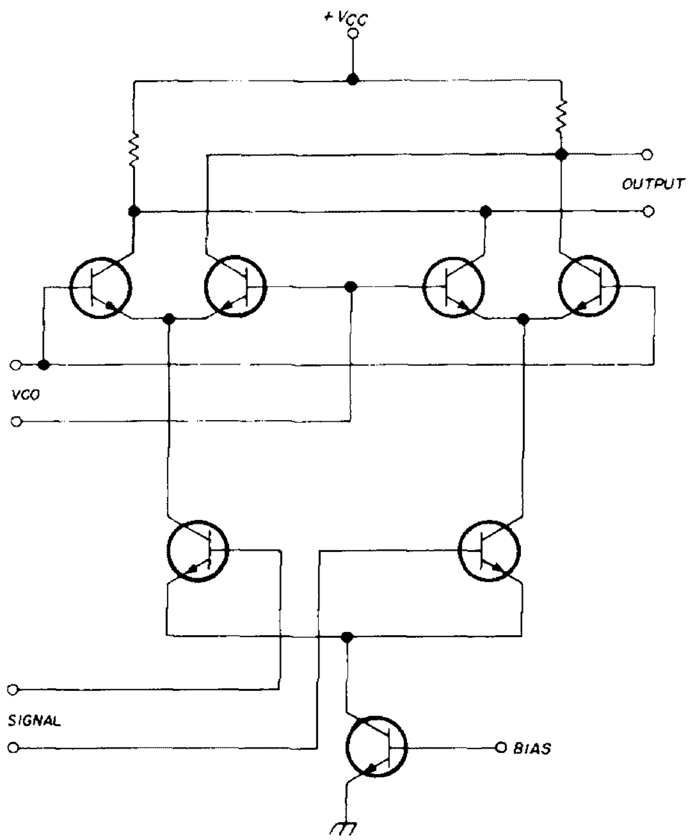


fig. 2. Basic phase comparator and multiplier circuits use differential transistor arrangement.

phase comparator

A simplified phase detector that is very similar to the one used in the NE561B is shown in fig. 2. This is the same differential amplifier I've described in previous *circuits and techniques*. Connect two differential amplifiers in a balanced arrangement and you come up with a balanced modulator, balanced demodulator or balanced mixer. The vco output is connected to the bases of the differential pairs; the signal is applied base-to-base of the emitter input transistors.

When the vco and input voltage are phase locked to the same frequency the output voltage is zero. An error voltage develops when the signal drifts off. In the

case of fm demodulation the recovered error voltage is actually the demodulated fm wave.

The a-m demodulator uses the same basic balanced differential arrangement as the phase comparator. The vco output is applied to the bases of the differential pairs; the rf a-m signal is introduced through the insert transistors in the emitter circuit. The phase comparator is also used in the a-m demodulation system to phase-lock the vco to the carrier frequency. The difference frequency out of the differential multiplier circuit (fig. 2) corresponds to the frequency separation between the sidebands and carrier of the a-m signal, hence the circuit reproduces the original audio modulation.

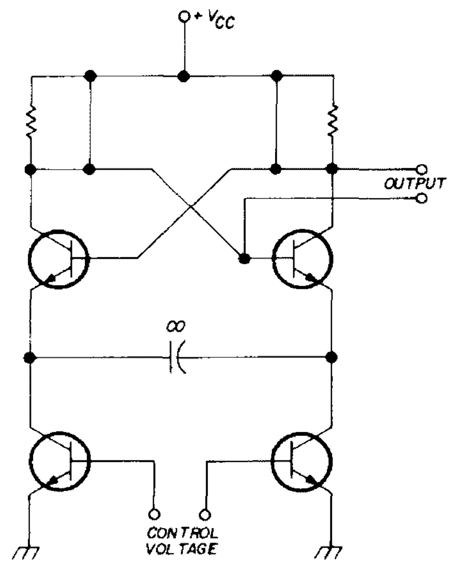


fig. 3. Voltage-controlled oscillator is a multivibrator. Basic frequency is controlled by capacitor connected between the emitters; fine tuning is provided by voltage that controls emitter current.

vco

The voltage-controlled oscillator is an emitter-coupled multivibrator similar to the circuit shown in **fig. 3.** The external timing or tuning capacitor is connected between the emitters of the multivibrator transistors. At any instant of time one of the transistors is turned on and the other is off. The timing capacitor is alternately charged and discharged by the voltage-controlled current source in the emitter circuit. The control voltage can be in the form of a dc error voltage from the phase comparator through the limiter, or an external control voltage may be introduced.

practical circuits

The circuit in fig. **4** shows how the Signetics NE561B may be used as an fm detector. The fm input signal is connected to pin 12 and 13; audio output appears across the 15k load resistor connected to pin 9. The capacitor between pin 10 and ground establishes the proper de-emphasis.

The capacitor between pins 2 and 3 determines the frequency of the voltage-controlled oscillator. Typical capacitor values for intermediate frequencies of 4.5 and 10.7 MHz are given in fig. **4.** For fine tuning a small capacitor may be placed in parallel with C_o (C_x in fig. **4**) or a small variable voltage may be applied to pin 6. This can be in the form of a potentiometer connected across the power supply. The low-pass loop filter components are connected between pins **14** and **15.** Component values, which depend on the i-f frequency, are shown.

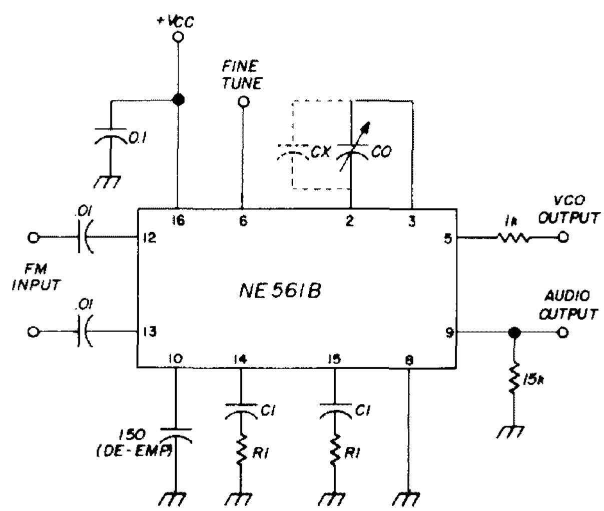


fig. 4. Use of the phase-lock loop IC as an fm detector.

The frequency range over which the vco will track the signal frequency can be

altered by applying appropriate dc control voltages to pin 7 from an external and adjustable supply voltage source. In some specialized application it may be necessary to offset an internal dc component. This can be done with an external voltage source applied to pin 11.

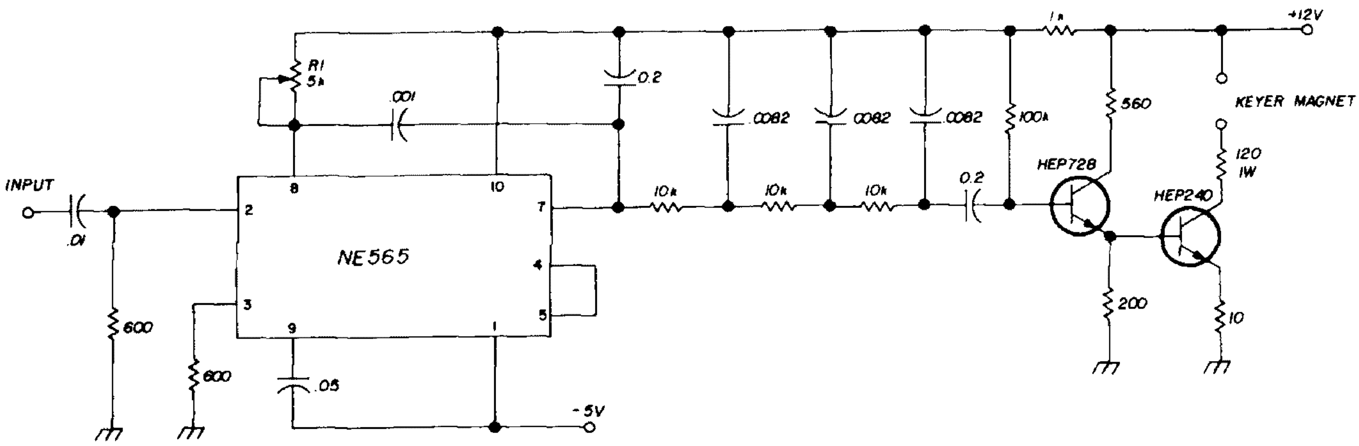


fig. 6. Phase-locked RTTY demodulator requires no bandpass filters. R1 is adjusted so internal vco frequency is at same frequency as input.

The same IC is setup for synchronous a-m detection in fig. 5. Note that the vco, phase discriminator and low-pass filter stages are active as they were for fm demodulation. However, the multiplier circuit is also necessary for synchronous a-m detection.

In the a-m detection process the phase-lock loop is locked to the signal carrier frequency and the vco output serves as the local oscillator signal. The appropriate quadrature relationship for synchronous demodulation is established by an external 90° RC phase-shift network. These components, R_AC_A and R_BC_B in fig. 5, must satisfy the time-constant formula

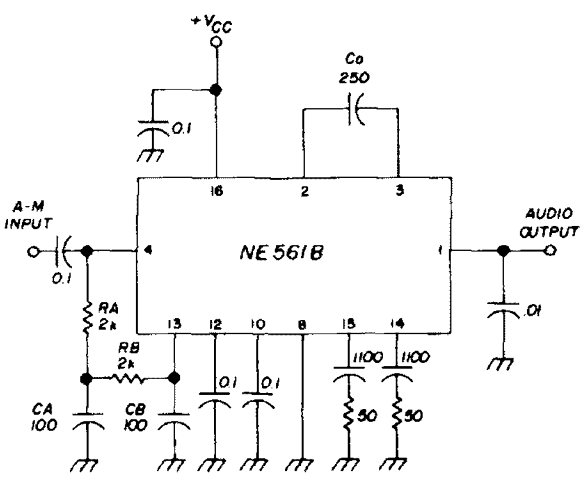


fig. 5. Synchronous a-m detector circuit. Component values shown here were selected for broadcast band (1.5 MHz).

$$R_AC_A = R_BC_B = \frac{1}{2\pi f_o}$$

Where f_o is the operating frequency. The values in fig. 5 were selected for the a-m broadcast band (1.5 MHz). In this circuit the 250-pF capacitor between pins 2 and 3 tunes the IC across the broadcast band; the low-pass filter components are suitable for broadcast-band operation.

A less complex integrated-circuit phase-locked loop such as the Signetics NE565 can be operated as an RTTY demodulator without the use of bandpass filters. This IC consists of a phase detector, amplifier and voltage-controlled oscillator. Range extends between a fraction of a cycle to 500 kHz.

In the fsk demodulator circuit in fig. 6 the incoming frequency-shift signal moves rapidly between the mark and space frequencies. As it does the phase-lock loop locks on to the input signal, moving from one input frequency to the other, with a corresponding dc voltage shift at the output. The simple RC filter across the output is designed to remove frequency components between the maximum keying rate and twice the maximum input frequency. The filter components in fig. 6 were chosen for amateur opera-

tion (100 wpm maximum) with a 455-kHz i-f strip.

multi-ic loops

An integrated-circuit phase-locked loop does not have to be confined to a

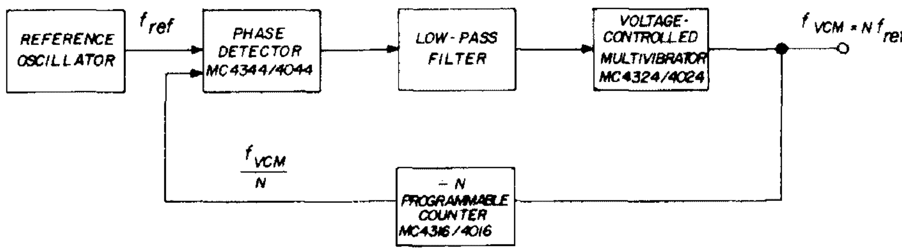


fig. 7. Frequency synthesizer uses programmable divide-by-N counter within the phase-locked loop. Typical Motorola ICs for this circuit are shown.

single IC. There are a variety of special-purpose ICs that can be combined into phase-lock systems and tailored for specific applications. For example phase-detector ICs such as the Motorola MC4344/4044 and voltage-controlled multivibrators such as the Motorola MC4324/4024 can be combined with a low-pass filter to form a simple phase-locked loop.

A system of separate integrated circuits is especially attractive for building complex circuits such as a frequency synthesizer. A typical arrangement is shown in **fig. 7.** With a programmable counter in the loop the system can be programmed to provide practically any output frequencies. The output signal is phase locked to a stable reference oscillator so output stability is essentially that of a crystal oscillator.

In the circuit of **fig. 7** the reference oscillator signal is compared to the signal from the programmable counter. The input to the programmable counter comes from a vco which is controlled by the error voltage from the phase detector. With a 1-MHz reference oscillator and a divide-by-2 counter, the output signal is 2 MHz. With a divide-by-3 counter the output is 3 MHz; divide-by-10 counter, 10-MHz output, etc. By placing a divide-by-N counter (where N is any number) in the loop you can build a frequency synthesizer which will provide any desired output frequency.[3]

army loop antenna — revisited

A simple method of tuning the ends of a small resonant-loop antenna has been evolved by James E. Taylor.[1 2] This technique, shown in **fig. 8,** eliminates lossy loading-coil arrangements. Maximum horizontal length for operation on 80 meters is 20 to 25 feet. Although Taylor used RG-8/U cable for his antenna any large diameter conductor is satisfactory.

The capacitive element in **fig. 8** consists of 300-ohm twin-lead. Note that the ends of the resonant loop connect to opposite conductors of the twin-lead. The loop is resonated by clipping even amounts off each end of the twin-lead. (Shortening the twin-lead lowers its capacitance, raising the frequency of the resonant loop.) The large capacitor across the feedline provides proper matching.

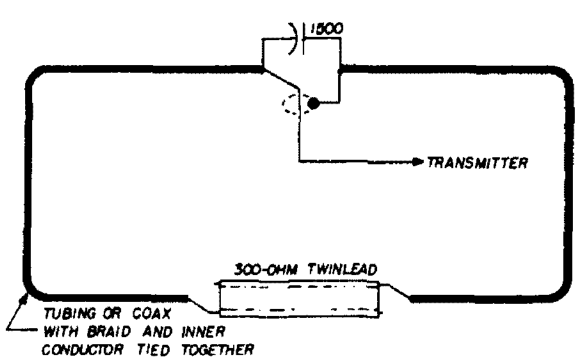


fig. 8. Balanced Taylor-loop antenna.

linear transistor amplifier

Despite the popularity of high-frequency ssb transmission there have been few bipolar transistors designed specifically for linear operation at higher power levels; most of the rf power transistors

has been designed for class-C operation. However, the new RCA 2N6093 is designed specifically for class-AB linear operation. A single 2N6093 can provide 75 watts PEP output at 30 MHz; a push-pull amplifier with two devices is capable of 150 watts PEP from 2 through 30 MHz.

To provide operation in class B or AB the bipolar transistor must be slightly forward biased. Unfortunately, this biasing increases the possibility of destroying the transistor because of secondary breakdown. However, the new 2N6093 transistor can be operated safely at the required dc operating levels. The device uses special subdividing of the emitter and an appropriate resistive ballast. This added resistance improves stabilization and permits low distortion operation.

Included within the transistor package is a temperature-sensing diode for bias compensation and run-away protection. This diode insures that the forward bias varies with temperature in the same manner as the base-emitter voltage of the transistor. An external current amplifier is required to build the diode sensing signal up to a usable level.

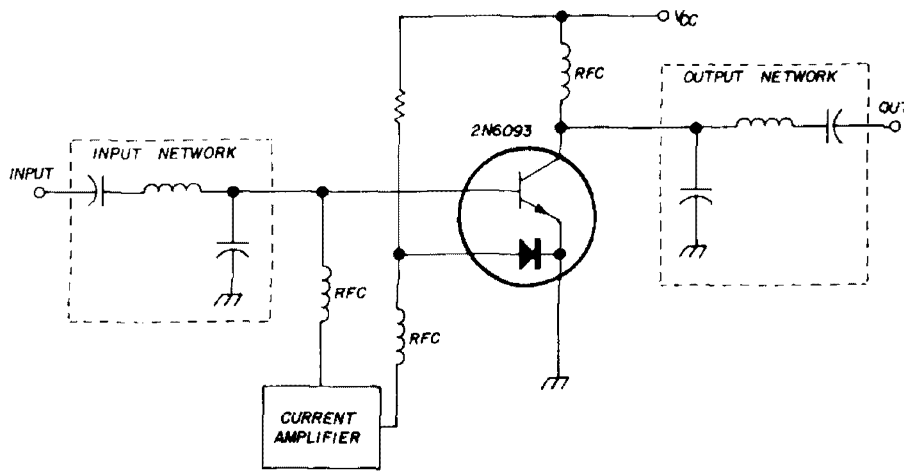


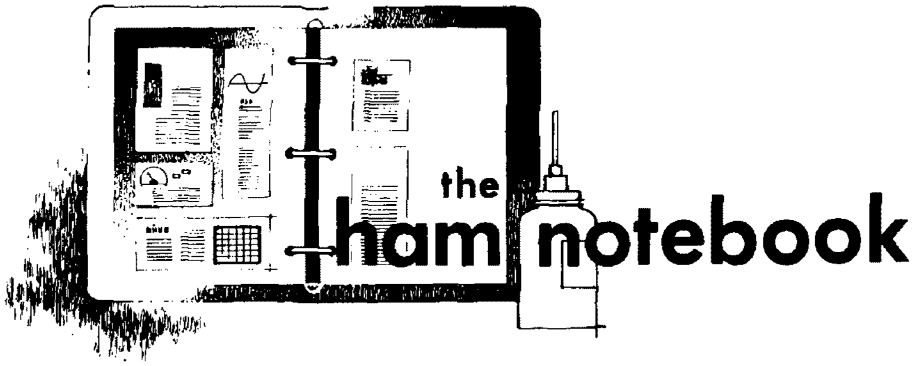
fig. 9. Basic rf linear amplifier circuit for new high-power linear rf transistor.

references

1. J. A. Koehler, VE5FP, "Phase Locked-Local Oscillator," *ham radio,* March, 1971, page 6.
2. John K. Webb, WØAHM, "A Synchronous Adapter for Communications Receivers," *CQ,* June, 1957, page 30.
3. L. W. Danz, WA5SKM, "Frequency Synthesis," *ham radio,* December, 1969, page 42.
4. F. M. Gardner, "Phase Lock Techniques," John Wiley and Sons, New York, 1966.
5. Larry Steckler, "One IC replaces entire i-f strip," *Radio-Electronics,* November, 1970, page 59.
6. "General Description of the NE560B/561B Phase Locked Loops," Signetics Applications Memo.*
7. J. Mattis and H. Camenzind, "A New phase-locked loop with high stability and accuracy (SE/NE565)," Signetics Applications Memo.
8. G. Nash, "Phase-Locked Loop design fundamentals," Motorola Applications Note AN-535, 1970.
9. E. Renschler and B. Welling, "An Integrated Circuit Phase-Locked Loop Digital Frequency Synthesizer," Motorola Applications Note AN-463, 1969.
10. J. DeLaune, "MTTL and MECL Avionics Digital Frequency Synthesizer," Motorola Applications Note AN532, 1970.
11. Alan B. Grebene, "The Monolithic Phase-Locked Loop – a Versatile Building Block," *IEEE Spectrum,* March, 1971, page 38.
12. James E. Taylor, "Twin-lead Tunes Foreshortened Balanced Loop Antenna," *Electronics,* April 12, 1971, page 88.
13. I. Chang, A. Locke, "Use of the RCA 2N6093 HF Power Transistor in Linear Applications," RCA Application Note AN-4591.

*Signetics Applications memos are available from Signetics Corporation, 811 East Arques Avenue, Sunnyvale, California 94086. Motorola Applications Notes are available from Motorola Semiconductor Products, Technical Information Center, Box 20912, Phoenix, Arizona 85036.

ham radio



relay activator

Occasionally, ssb exciter or transceiver relay contacts give trouble when they turn on a relay in a linear amplifier. This is part of the vox or push-to-talk system; any hang-up can be very irritating.

Some older ideas like placing an 0.5 μF capacitor in series with possibly 470 ohms, across the relay line between the exciter and the amplifier, may help. However, the cost of doing the job with a transistor is small, leaving the relay contacts with little current to handle.

If your amplifier is similar to the Henry 2K, which provides 14 to 18 volts to its relay, and requires less than 300 mA to activate it, this method will work.

Fig. 1 shows the simple circuit. The pnp transistor can be replaced with an npn if the linear provides negative instead of positive voltage to its relay phono jack. Some problems were encountered in locating a suitable transistor that would operate by shorting its base, through a resistor, to ground, in order to activate the amplifier relay, without tending to have the collector current creep toward a destructive value. However, the little 2N4918 30-watt pnp from Motorola ($1.65) does the job very well. It has a hole permitting it to be fastened to the inside of the minibox as a heat sink – but does not seem to get warm. Since the collector is grounded no insulation is needed between the 2N4918 and the box. This device is rated at some 40 volts; for other voltages, see adjacent type numbers (2N919, 60 volts; 2N4920, 80 volts).

The addition of a 1N2070 diode, with 400 volts inverse breakdown rating, provides protection from inductive surges from the relay coil. It was not installed initially; no problems were encountered without it. Presumably, some handy diode will be available to use as a substitute, if desired.

The base resistor worked with up to 5.6k ohms but, above that, the amplifier relay did not get enough current to pull in. Values as low as 1.5k ohms performed equally well but with slightly higher base current to be handled by the exciter relay contacts. There was no sign of creeping collector current. Using 4.2k ohms, the transistor handled 300 mA for the amplifier relay, and required only 8 mA on the contacts of the exciter relay.

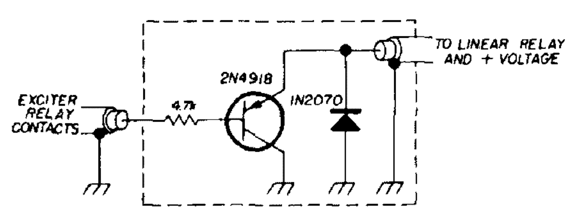


fig. 1. Relay activator operates linear amplifier relay with little current on exciter relay contacts.

Much higher current than 300 mA probably can be handled by the 2N4918,

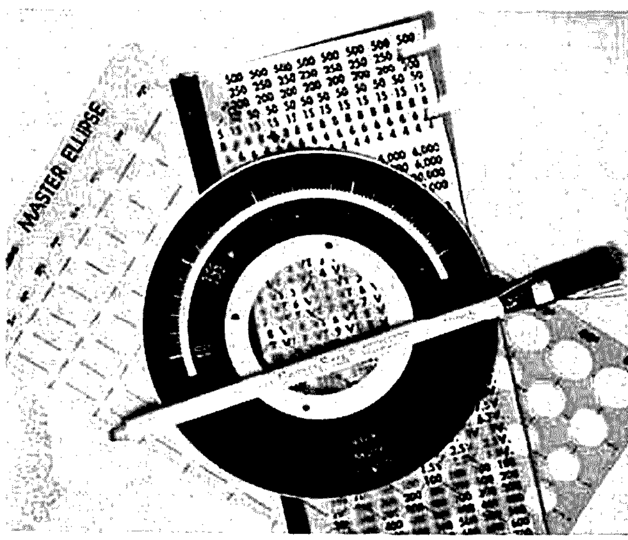
in which case the base resistor can be increased in order to limit the base current which will have to be handled by the exciter relay contacts. Similarly, some Henry 2K relays require only 250 mA; any relay coil requiring less than 300 mA will require a smaller base resistor for satisfactory operation.

Some interesting variations of the circuit in fig. 1 may be possible if an additional driving transistor is tolerated.

Bill Conklin, K6KA

putting the HW-12 on MARS

The Heathkit HW-12A ssb transceiver can be modified easily for use on the 4-MHz Navy MARS frequencies without any electrical changes. Operation on Navy MARS requires a frequency range up to about 4050 kHz. In the HW-12A this is accomplished by changing the tuning range from the original 3800 to 4000 kHz to 3850 to 4050 kHz. This is done by re-adjusting vfo coil, L5, and the vfo trimmer capacitor so the vfo tunes from 1.5433 to 1.7433 MHz (original vfo frequency range is 1.4933 to 1.6967 MHz).



HW-12A dial-conversion kit.


Heathkit HW-12A with new MARS dial.

To improve the vfo output level across the new frequency range the driver grid coil, L2 (40-516) and the driver plate coil, L3 (40-513) must be unsealed. Peak the top slug of L2 at 4000 kHz and the bottom slug of L2 at 3900 kHz; L3 is peaked at 3950 kHz. Depending on component tolerance, you may have to change one of the two fixed-value power-amplifier tuning capacitors (C206 and C207) from 68 to 47 pF.

As a final touch, to maintain appearance and provide direct frequency readout, the vfo dial is modified. A very professional appearance can be obtained by erasing the existing dial numbers with an ink eraser, using an elipse template as an erasing guide. A fine-tipped eraser pencil works best (see photo). The black areas are easily removed from the white plastic dial. Black decals are then applied inside the erased areas at 3850, 3900, 3950, 4000 and 4050. A touch of clear plastic model paint over the decals completes the job.

I have had excellent results throughout the 4th Naval District with the modified HW-12A and a dipole antenna. The dipole consists of approximately 108 feet number-16 insulated wire, twisted around nylon sash cord for strength. It is center-fed with a 1:1 balun and RG-8/U coaxial cable; vswr at 4 MHz is 1.15:1.

David M. Stahley, K8AUH

crystal-controlled frequency markers

A simple frequency marker for your sweep generator can be built from several surplus crystals. Piezoelectric crystals function as highly-selective notch filters when placed in shunt with an rf signal, and have been used successfully in the grid and plate circuits of fm repeater receivers to short unwanted frequency components to ground.

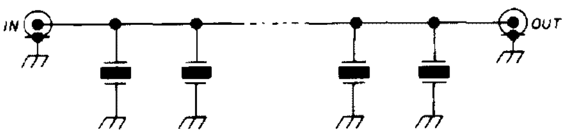


fig. 2. Simple crystal-controlled frequency marker uses shunting characteristics of quartz crystals.

The crystal frequency marker unit in **fig. 2** consists of several parallel-connected crystals mounted in a minibox. The output of the sweep generator is passed through the device to the circuit under test. Since energy at the resonant frequency of each crystal is shunted to ground, there is a suck-out for each of the crystals used in the circuit. In addition, if the crystals are overtone types, frequency markers are also shown at the harmonic frequencies.

The small amount of capacitance contributed by the circuit in **fig. 1** tends to shunt the rf signal more and more as the sweep signal increases in frequency. This appears as a slight negative slope on the oscilloscope display.

Earnest A Franke, WA4ADK

high-voltage step-start circuit

Fig. 3 shows a fail-safe step-start circuit for the primary of high-voltage power supplies that is simple and effective, having been used for many years in various applications. With the present trend to higher and higher capacitances in power supplies this start circuit it a simple precaution to extend the life of diodes and filter capacitors by reducing the initial surge charging current.

Furthermore, in the case of an overload, which may not react to other protective circuits, this step-start circuit will revert to its initial state, thereby providing a measure of protection. With appropriate connections this circuit may be used with either 120 or 240 volts ac.

Relay K1 is the contacter normally used to energize the high-voltage primary winding. As it closes a voltage drop will occur across R1. This voltage drop is adjusted to suit the requirements of the power supply. The voltage drop across R1 diminishes after the initial surge, and relay K2 is energized, shorting out R1. On a 240-volt system K2 may be connected to the center-tap of the power transformer, or to the neutral.

Marv Gonsior, W6VFR

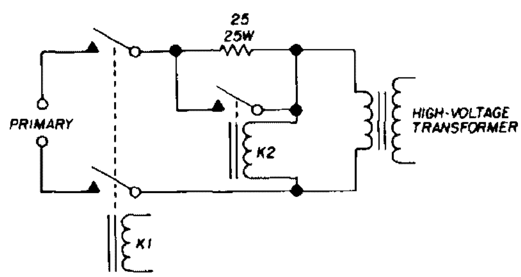


fig. 3. High-voltage step-start circuit.

rotator improvement

If you have an Alliance T-45 rotator (or similar non-automatic rotator) a simple circuit modification will eliminate the tiresome task of holding down the control bar with your finger. Just install a dpdt switch, with center off, in parallel with the contacts on the control-box switch. Switch mounting is a bit of a problem since there isn't room enough in the T-45 control-box to accommodate the new dpdt switch. For my control box I put the added switch in a small metal file box complete with direction labels (clockwise and counter-clockwise). Now I have one hand free for tuning the rig while I run the antenna around.

Ed Mitchell, WAØVAM

new products

ST-6 RTTY demodulator

HAL Devices now offers a standard wired ST-6 RTTY Terminal Unit. The unit is enclosed in a very attractive cabinet with all control and connector functions clearly labeled. Because of the standardization required to produce the silk-screen from which all front and real panels are labeled and drilled, wired units are available *only* in the maximum feature configuration for both shifts with autostart and anti-space. An additional feature of all wired units is the inclusion of meter-switching to allow use of the meter for both a tuning indication and to indicate loop current.

The cabinets are 3½" high x 17" wide x 12" deep and are available in rack-mounted or table-top configurations. The top and bottom panels of the cabinet are easily removable for access to the circuits boards and other components. These units are wired by skilled professional personnel, and all tuned circuits are aligned with an electronic counter to a tolerance of 5 Hz for the 850-Hz shift circuits and 3 Hz for the 170-Hz circuits. All adjustments are pre-set at the factory and all modes and features of each unit are tested in actual operation, receiving and printing RTTY signals.

All wired HAL Devices products carry a 1 year warranty against defects in materials and workmanship. The wired ST-6 is priced at $280.00 plus postage; shipping weight is approximately 23 pounds. When ordering be sure to specify which cabinet style, rack or table-top, is preferred and include sufficient remittance to cover postage and insurance. Limited staff time at HAL Devices as well as the great care (and therefore time) taken in the construction of each unit means that delivery times will be long, and at least 1 month should be allowed for each wired ST-6 ordered. Wired units are constructed on an as-received basis, the earliest orders being processed and constructed first. HAL Devices will acknowledge receipt of all orders for wired units and will indicate the projected delivery date at that time.

HAL Devices also offers ST-6 parts kits for the amateur who wants to build his own. To order, write to HAL Devices, Post Office Box 365, Urbana, Illinois 61801. For more information, use *check-off* on page 94.

digital frequency meter

The Micro-Z FM-36 frequency meter includes all of the features that have made the FM-6 so popular — high quality and accuracy, ease of assembly, guaranteed operation, small size and cost — plus the added convenience of a third digit to simplify switching and readout.

More Details? CHECK—OFF Page 94

This instrument is useful in the service shop for calibrating audio and rf signal generators or testing transmitter and receiver oscillators and multipliers. Simply connect the FM-36 to the output of your transmitter with the coax T-connector supplied, and your frequency is measured and displayed whenever you are on the air. In addition, a companion prescaler is available for vhf operation.

The FM-36 operates from audio frequencies through 35 MHz with input signals from 0.4 to over 200 volts. The circuit is completely self-contained with a 117-Vac power supply. The FM-36, factory assembled and tested, is $164.50. The FM-36 kit, including all parts and detailed assembly instructions, is $134.50 from Micro-Z Electronic Systems, Box 2426, Rolling Hills, California 90274. For more information use *check-off* on page 94.

cordless electric soldering iron

The new cordless electric soldering iron offered by the Technical Equipment Company provides an output of 40 watts with a heavy-duty 10 amp-hour NiCad battery. The Express 2000 features instant heat, negligible magnetic field, two interchangeable tips (25- and 40-watt), complete portability, and is solid-state-device proof — battery operation assures complete electrical isolation.

The Express 2000 cordless soldering iron has an operating voltage of 1.25 Vdc. This is furnished by the rechargeable battery. Each single charge of the battery provides 120 soldering operations; a full recharge takes 14 hours. The useful life of the battery is 3000 recharge operations. The soldering tips are chrome nickel steel with pure nickel points. Accessories include a 110 Vac charger, 110/220 Vac charger and recharge case with battery. For more information use *check-off* on page 94, or write to Technical Equipment Company, Post Office Box 247, Bothell, Washington 98011.

fiberglass mobile antennas

The new Shakespeare fiberglass antennas are a breakthrough in the design of mobile radio antennas. Two models are available: the style 456 for low-band, 25 to 57 MHz; and the style 455 for high-band, 144-174 MHz. Every effort has been made by the Shakespeare engineers to design an antenna that delivers the best electrical and mechanical performance without any compromises.

Particular care was given to eliminating built-in losses that absorb rf energy. This was accomplished with a low-dielectric taper-molded helix that offers precision repeatability, silver-plated solid-copper inductor to minimize coil heating, stranded-silver-conductor whip for large current-flow area, and separation of body metal from coil circuitry to reduce stray capacitance and leakage currents. These features combine to offer a high-Q circuit for maximum performance.

The flexible fiberglass tip inserts directly into the lower housing and eliminates the need for a base spring. The pressure-molded housing permanently seals the coil against the outside environment. A set-screw adjustment is provided so the tip can be slipped in or out for lowest swr. Adapters are available to adapt the style 455 and 456 to other popular brand base mounts.

The style 456 is a 1/4-wave antenna with a power rating of 100 watts. Standing-wave ratio is 1.3:1 or less when cut to exact frequency; input impedance is 50 ohms. Bandwidth is 300 kHz at 25 MHz, increasing to 1 MHz at 50 MHz.

The style 455 is a 5/8-wave antenna, power rated at 100 watts, with a power gain of 2.7 dB. Standing-wave ratio is 1.3:1 or less when cut to exact frequency; input impedance is 50 ohms. Bandwidth is ±2.5 MHz for 2:1 swr. Frequency range is 144 to 174 MHz. For more information, use *check-off* on page 94, or write to Shakespeare, Post Office Drawer 246, Columbia, South Carolina 29202.

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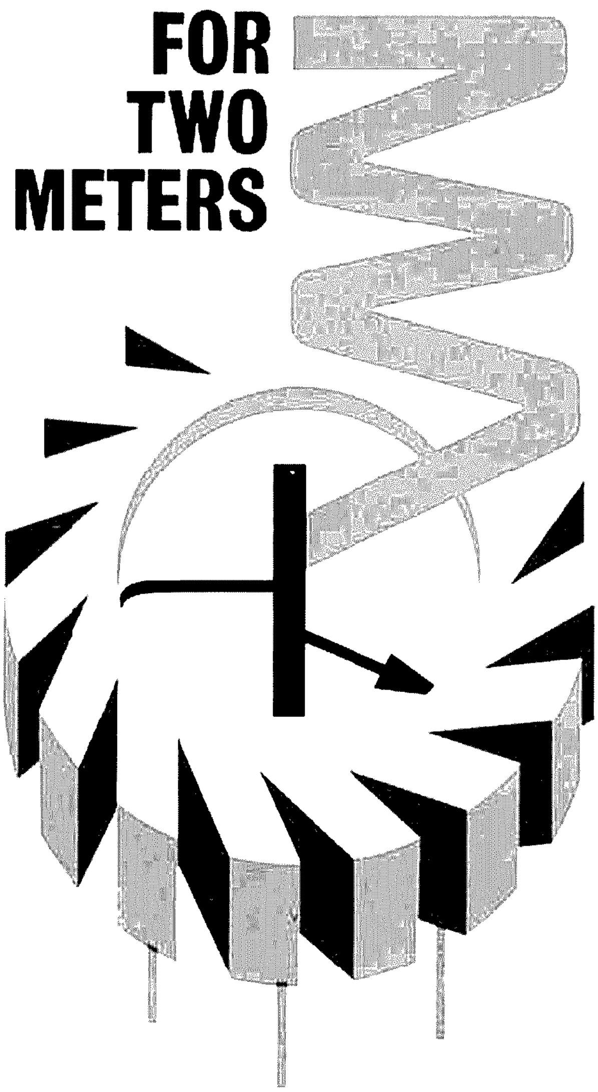

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ham radio magazine

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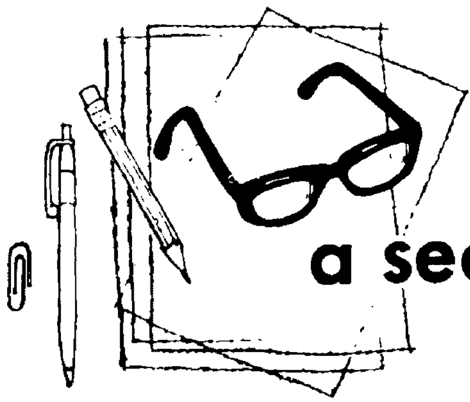


a second look

by jim fisk

Semiconductor microwave power devices — diodes that use drift time to generate large amounts of rf power — are finding their way into more and more commercial microwave equipment. Although these devices were predicted theoretically nearly fifteen years ago, it wasn't until 1963 that a practical solid-state power-generating device was actually built.

The Gunn diode or bulk-effect device is a simple chunk of n-type gallium arsenide that generates microwave power directly when a voltage is placed across it. When a constant voltage is applied to the semiconductor material the current through it fluctuates at an extremely rapid rate, although somewhat randomly. If the slice of semiconductor is less than about 0.005-inch thick the current no longer fluctuates randomly, but rises and falls in a cyclic way, generating microwave power.

The bulk-effect device is inherently broadband and frequency output is determined by the circuit in which it is used. Present maximum power output in the CW mode is 1 watt at 5000 MHz and pulse powers of 2000 watts at this frequency have been obtained in the lab.

Avalanche diodes are another source of microwave power. These diodes operate in two basic modes: Impatt and trapatt. Impatt (for IMPact Avalanche Transit Time) oscillators use the negative resistance that results from a combination of internal secondary emission and bunched current carriers that drift through the solid-state material and deliver rf power by causing external circuit current which is 180° out of phase with the applied voltage.

In the Trapatt mode (for TRApped Plasma Avalanche Triggered Transit) the diode operates in the Impatt mode at some high microwave frequency and the desired output is taken at a subharmonic of the Impatt frequency. Trapatt diodes can deliver 10 watts CW at 1000 MHz with about 60% efficiency. The much less efficient Impatt diodes operate at about 10% efficiency, but at much higher frequencies — up to 150 GHz.

Although the cost of off-the-shelf devices is still fairly high, some enterprising engineers have discovered that the simple rectifier diodes in your amateur gear may be uhf Trapatt oscillators in disguise. They found that ordinary Fairchild FD-300 rectifier diodes yielded 68-watt pulses at 630 MHz! Out of 100 FD-300s purchased for the experiment, 83 oscillated with greater than 35-watts *average* output. When three diodes were mounted in parallel they provided repeatable 395-watt pulses at 570 MHz; efficiency was a surprising 75%.

The Fairchild FD-333 rectifier, which has higher capacitance and greater breakdown voltage than the FD-300, has been made to generate a respectable 152-watt peak power at 630 MHz, so it's doubtful that these are isolated cases. There must be many other common silicon rectifiers that perform as well or better as microwave generators. If you are already using commercial microwave diodes in your uhf equipment, or have found other low-cost rectifiers that yield useful amounts of uhf power, I would like to hear about it.

Jim Fisk, W1DTY
editor

publisher's log

skip tenney
W1NLB

Who won the Signal/One-Alpha 70 package and the many other fine prizes in *ham radio's* 1971 Sweepstakes? This question has been asked rather frequently of late since many readers failed to see the announcement of the top-prize winners in "A Second Look" in our August issue.

prize winners

Here are the complete results of the Sweepstakes:

grand prize winner
Signal/One CX-7
Alpha 70 Linear by ETO
Mr. John F. Longley, W2ANB

second prize winners
two-Meter solid-state transceivers
W2CNB
W4YPC
WB8DUO

The big moment finally arrives as Ham Radio's advertising manager, Hilda Wetherbee, draws the winning Sweepstakes entry.

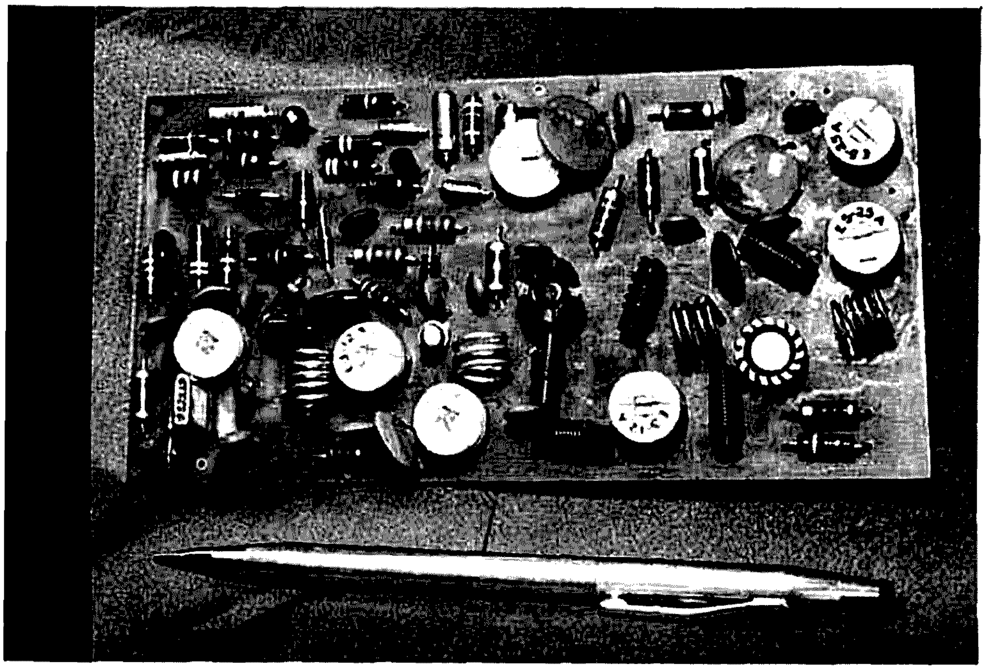
third prize winners

WØEAD	WØEZE	K7HNV
W5PQA	WN5ZGC	W2ESH
WN1MRY	WB4OEX	KØGZW
WN1ONM	K5FRI	KØARK
K8OSR	K9WTT	WN8HPO
W7HJR	WA4AUC	WA8VSJ
W7HJY	WN4PFR	WA8SXA
K7ZVA	WB4LOY	W8LPY
WA5PMY	WA4YTK	WA8ZJL
WAØTRF	W2WKL	WN4QYG
W1PEX	W4BRE	W6DZK
K9TTE	W4FES	WN6OBJ
WA7GUV	WA4FIG	K6BOY
WAØNEH	W4ZSK	W6JN
W4ICL	K4OG	WN6LAJ
WN5YBM	WAØQJK	WA6MQT
W9UCI	WN1KQB	

On behalf of all of us here at *ham radio*, we want to thank the many thousands of you who entered for your participation. Our only regret is that everyone could not be a winner — maybe you'll be the lucky one next time. Let's hope so.

We are quite proud of this issue of *ham radio*. A check of our advertising index will show more advertisers in this issue than have appeared in any issue of any amateur magazine during 1971 (and probably for a good while before that). Although this is the result of a lot of hard work by our staff, it is also very much a credit to you, our reader. Your support of our advertisers has made *ham radio* a good investment for them. Keep up the good work!

Skip Tenney, W1NLB
Publisher



sonobaby

vhf fm transmitter

A miniature, solid-state design that provides low cost through the use of readily available surplus Sonobouy components

Robert Shriner, WAØUZO, Robert Heptig, KØPHF and Earl Gill, WBØEDA

Although interest in two-meter fm has been increasing at an explosive rate, there have been few practical circuits published for high-performance vhf-fm transmitters. The solid state fm transmitter described in this article is designed for operation from 12-volt dc power supplies, provides 2-watts rf output and has excellent audio quality. Best of all, it may be duplicated easily and at low cost.

The advent of surplus Sonobouy transmitters,* and the ingenuity of some enterprising amateurs who discovered how easily they could be put on 2 meters, have resulted in a high quality 2-meter fm transmitter that is ideal for use in an fm base station, mobile unit or walkie-talkie. The 2-watt rf output has been more than adequate for working our repeater from 35 miles away using nothing but a ¼-wave ground-plane antenna.

The original Sonobouy transmitter consists of two 4 x 12-inch printed-circuit boards (shown in photo) and is quite large and cumbersome. We decided to see what could be done about civilizing this

*Surplus Sonobouy transmitters are $14.95 postpaid from Monks Electronics, 313 Old Farms Road, Simsbury, Connecticut 06070.

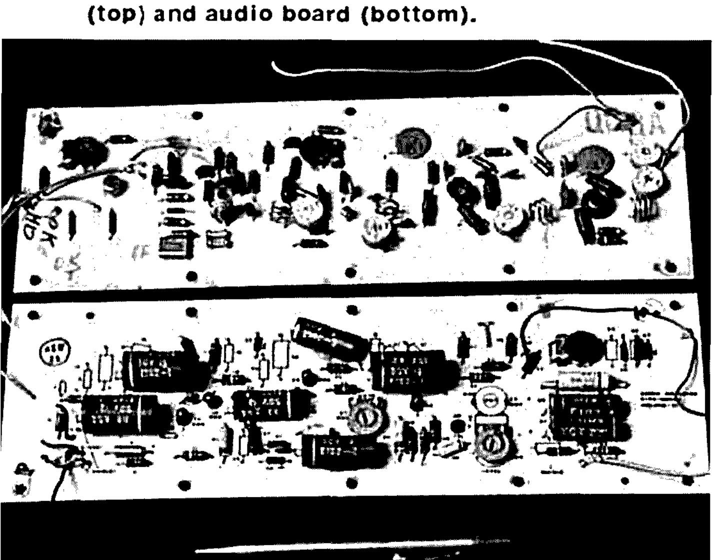
brute and reducing its size with a minimum of work and cost.

The audio section was the first to go; it simply had too many parts for the job it had to do. We decided to replace it with a simple two-stage circuit that had about the right amount of gain for 5-kHz deviation (see fig. 1). The new audio section uses an inexpensive field-effect input transistor (Q1) plus one of the transistors from the original Sonobouy audio board (Q2). With this circuit, deviation is controlled quite effectively by potentiometer R1. (Additional deviation may be obtained by increasing the value of C5 to 10 μF.)

The original Sonobouy rf board appeared to use a normal amount of components and worked very well. However, the oscillator wasn't very stable after lowering the frequency from 20 to 18 MHz, so we used a new oscillator circuit that was based on proven designs. The new oscillator circuit is very stable and requires no adjustments or peaking—simply install a crystal.

The new 3 x 6-inch printed-circuit board shown in the photographs accommodates both the rf section and the audio section. This layout is the result of considerable parts shifting and several earlier boards; however, signal quality actually exceeds the quality of the original. This transmitter, which we call the Sonobaby, has proven so popular that a parts list was developed that would allow amateurs not having access to a Sonobouy to build one from scratch.*

Original Sonobouy transmitter board (top) and audio board (bottom).

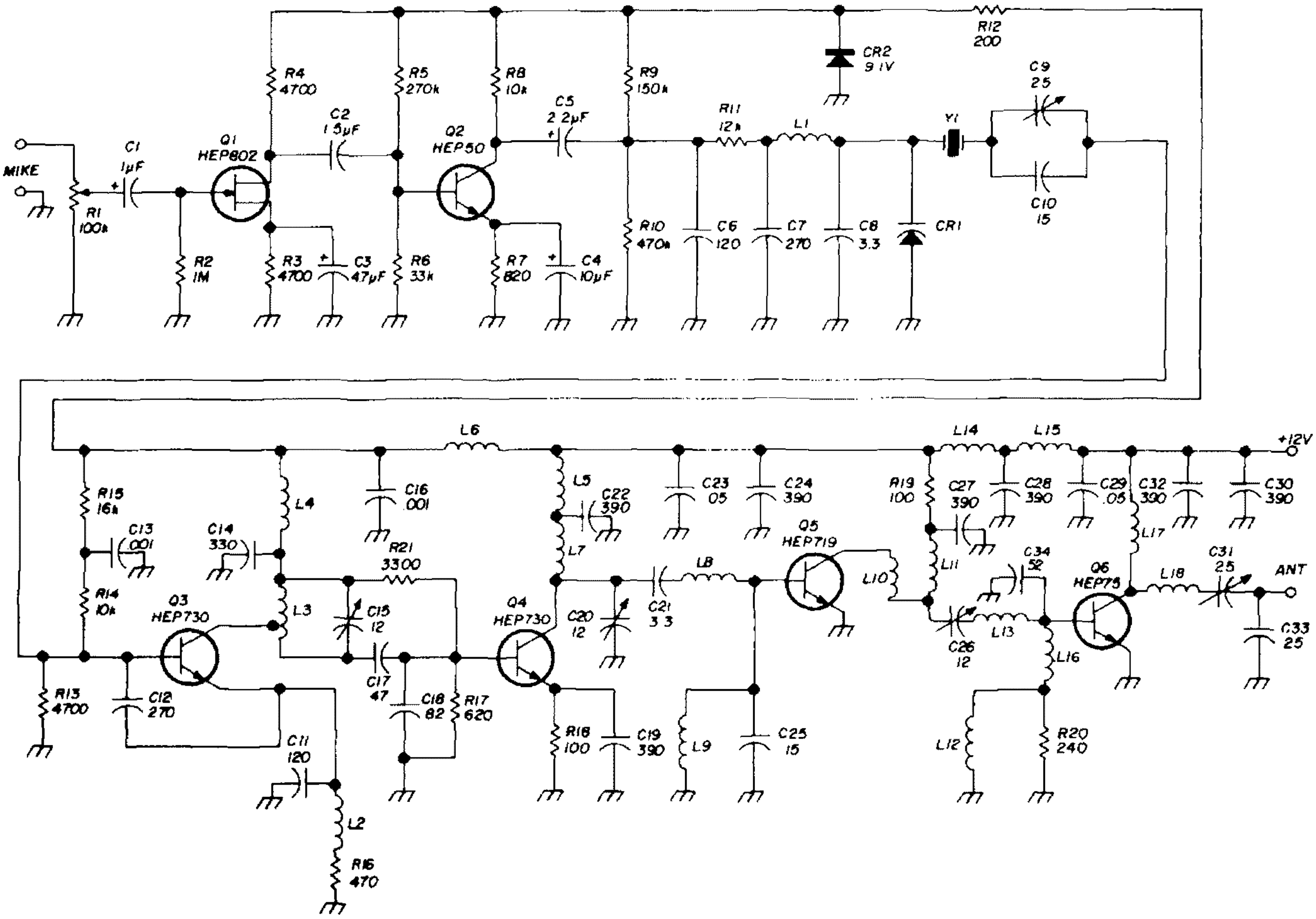


This Bearcat fm scanner monitor has been converted to a transceiver by installing a Sonobaby fm transmitter inside. Many vhf-fm receivers have been used in this manner to build complete transceivers and walkie-talkies.

*An etched and drilled epoxy circuit board can be purchased for $7.50 from Sonobaby, Post Office Box 92, Pueblo, Colorado 81002. A complete parts list, plus changeover information from the original Sonobouy transmitter and tuneup procedure is included with each board.

A semiconductor package for the Sonobaby vhf fm transmitter is available at a special price from Circuit Specialists. The transmitter package, consisting of Q1, HEP802, Q2, HEP50, Q3 and Q4, HEP730, Q5, HEP719, Q6, HEP75, CR1, MV2101, and CR2, HEP104, is priced at $9.05. A semiconductor package for the power supply, consisting of Q1, HEP624, CR-CR4, HEP175, and CR5, HEP605, is $5.00 postpaid. Order from Circuit Specialists Company, Box 3047, Spottsdale, Arizona 85257.

A complete kit of parts for the Sonobaby, including circuit board and all components (less power supply) is priced at $47.50 complete, plus shipping, from HAL Devices, Post Office Box 365H, Urbana, Illinois 61801. This parts kit includes an rf detector and 1-mA meter for use in tuneup.



C1	1 μF electrolytic (Mallory TAC 105M020P02)
C2	1.5 μF electrolytic (Mallory TAC 155M020P02)
C3	4.7 μF electrolytic (GE 76 F-02KC4R7)
C4	10 μF electrolytic (GE 76 F-02FC100)
C5	2.2 μF electrolytic (Mallory TAC 225M020P02)
C6,C11	120 pF Disc (Sprague 5GA-T12)
C7,C12	270 pF Disc (Sprague 5GA-T27)
C8,C21	3.3 pF npo Disc (Sprague 10TCC-V33)
C9,C31,C33	5/25 pF npo trimmer (Erie 557-000A-5-25)
C10	15 pF Disc (Sprague 5GA-Q15)
C13,C16	.001 μF Disc (Sprague 5GA-D10)
C14	330-pF Disc (Sprague 5GA-T33)
C15,C20, C26	3-12 pF npo trimmer (Erie 557-000A-3-12)
C17	47 pF Disc (Sprague 5GA-Q47)
C18	82 pF Disc (Sprague 5GA-Q82)

fig. 1. Schematic diagram of 2-watt vhf-fm transmitter.

the circuit

In the circuit in **fig. 1** a field-effect transistor, Q1, is used as an audio preamplifier. Transistor Q2 further amplifies the audio signal and passes it on to the varactor diode, CR1. The capacitance of CR1 changes with the audio signal, varying the frequency of the crystal-controlled oscillator, Q3. The zener diode, CR2, regulates the voltage to the audio amplifier and provides constant bias to the varactor.

The tank circuit of the 18-MHz crystal-controlled oscillator stage is tuned to 73 MHz; transistor Q4 doubles this signal to 146 MHz. Transistor Q5 drives the power amplifier, Q6, to 2 watts output on 2 meters.

C19,C22, C24,C27, C28,C30,C32	390 pF Disc (Sprague 5GA-T39)
C23,C29	0.05 μF Disc (Centralab DD503)
C25	15 pF npo Disc (Sprague 10TCC-Q15)
C34	52 pF Disc
CR1	MV 2101 varactor (Motorola) or Eastron VC2101
CR2	9.1 V zener (Motorola HEP104)
Y1	Crystal, order from International Crystal Company for Westinghouse Air Brake. Carry Phone II 20TS-1 Transmitter; specify operating frequency and commercial Standard
L1	15 μH choke (J. W. Miller 9310-40)
L2,L12	10 μH choke (J. W. Miller 9310-36)
L3	5 turns no. 16, 15/16" ID tapped at 1 1/2 turns
L4,L5,L14 L15	1.5 μH (J. W. Miller 9310-16)
L6,L9	4.7 μH (J. W. Miller 9310-28)
L7	3 turns no. 16, 15/16" ID
L8	7 turns no. 22, closewound on 3/16" plastic rod
L10	7 1/2 turns no. 22, closewound on 3/16" plastic rod
L11	5 turns no. 22 on 3/16" plastic rod, 5/16" long
L13	4 turns no. 16
L16,L17	12 turns no. 22, closewound on 3/16" plastic rod
L18	5 turns no. 16, 5/16" ID

construction

To build the Sonobaby from the original Sonobouy, simply remove the components from the original circuit boards and install them on the new Sonobaby circuit board. A few small parts must be purchased; notably the field-effect transistor and the 9.1-volt zener diode. It's a good idea to unsolder the transistors and the varactor before removing the other parts; this prevents excess heat from damaging them. Tag each of the semiconductors so you know what they are, and put them on a shelf out of the way as they should be the last parts you install on the new board.

For greater efficiency and power output rewind the coils as noted in the parts list. If you carefully examine the existing Sonobouy coils you will find that several of the original coils have the correct number of turns if they are moved to another location.

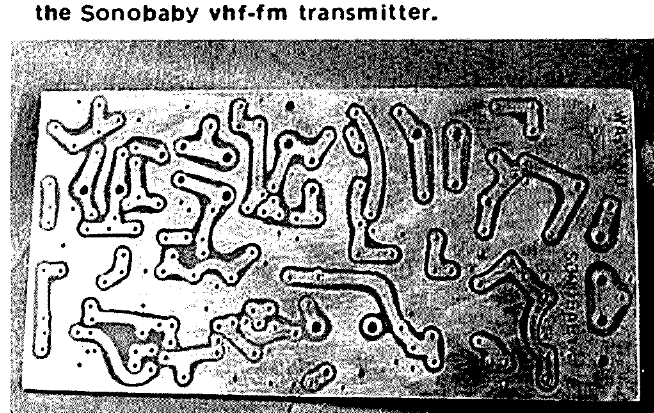
The coils wound on plastic rod stock are most easily made by first drilling two holes in the plastic rod the correct distance apart. The coils are then wound, passing the ends of the wire through the holes.

After all the components have been installed on the new printed-circuit board inspect it thoroughly for bad solder joints and short circuits before applying power. Connect some sort of indicating dummy load across the output (a wattmeter or a number-47 bulb), put a 500-mA meter in the 12-volt supply line and apply power. If you made no mistakes the unit will draw approximately 20 mA untuned; if it goes up in smoke you didn't inspect it carefully enough after installing the parts on the new circuit board.

tune up

When the transmitter has passed the first smoke test you can proceed with the tuning. First, tune C15 for maximum

Foil side of the printed-circuit board for the Sonobaby vhf-fm transmitter.



current through the 500-mA meter. Then, with a wavemeter (or grid-dipper in the diode position) tuned to 73 MHz and coupled to L3, adjust C15 maximum indication.

low to be measurable, is estimated at 10 milliwatts.

To put the Sonobaby on the air, connect a crystal or ceramic microphone to the input and adjust R1 for the best

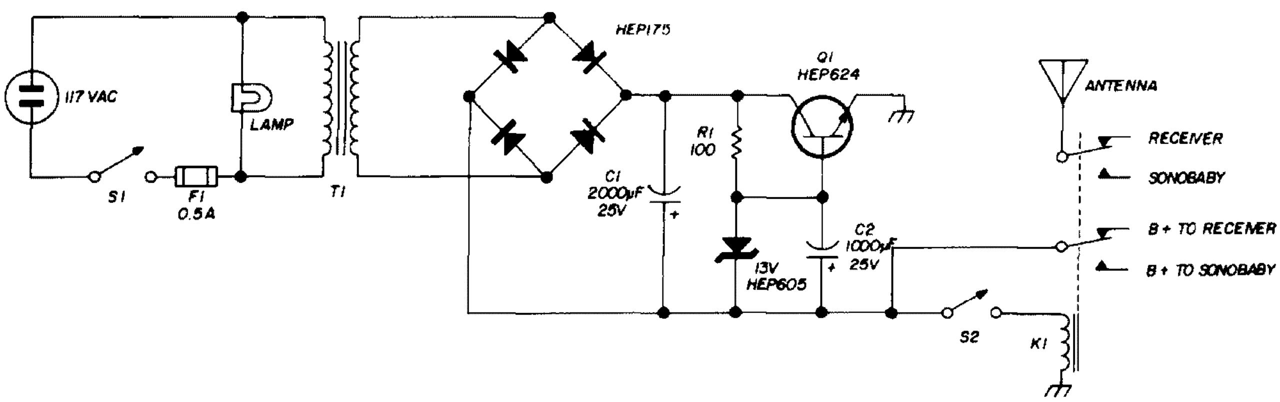


fig. 2. Power supply for the vhf-fm transmitter. Transformer T1 is a 12.6-volt filament transformer rated at 1 amp. Dpdt relay, K1, has 100-ohm coil.

The remainder of the alignment procedure is very straight forward: tune C20, C26, C31 and C33 in sequence for maximum rf output. Keep a check on the transistors during tuneup to make sure they don't get too hot. It is normal for them to be warm to the touch; if you can fry eggs on them, shut the power off then let them cool off before proceeding.

Capacitor C9 can be adjusted to put you precisely on frequency. If you have difficulty getting on frequency the value of C10 can be varied or the bias voltage set up by R9 and R19 can be changed. For best operation the bias voltage should be from +4 to +6 volts.

For the final step, couple an indicating wavemeter to the output and retune all capacitors for maximum indication. Check the second harmonic (292 MHz) to make sure it is well suppressed; if it is not touch up the tuning capacitors.

In the interest of good amateur practice (and conserving electricity) the value of R18 can be increased to as much as 1000 ohms for reduced power output. WAØUZO, who happens to live close to the repeater, has operated a Sonobaby with the final transistor removed and the antenna connected to L16. The circuit only draws 50mA, and output power, too

sounding audio. To observe the audio signal connect an oscilloscope to the collector of Q2.

transistor substitution

The bias resistors and capacitors shown in the schematic are correct for transistors salvaged from a Sonobouy transmitter. However, any substitution of transistors will likely require bias adjustments on that particular circuit.

Bias adjustment is very easy with a resistor substitution box. The procedure is simple: start at the crystal oscillator stage and adjust the bias resistors until the oscillator is putting out the desired signal; then move one stage at a time toward the final, adjusting each for maximum output.

After each change in capacitor value check for spurious and harmonic outputs. In some cases you may notice that a capacitor can be completely removed from the circuit for increased power output. Use extreme care – all capacitors in the circuit have a job to do; removing one is asking for trouble.

For example, if the 52-pF capacitor from the base of the final transistor to ground is removed, a wattmeter will indicate a decided increase in power

output. However, a careful analysis of the output signal will show that the power is all in spurious signals and harmonics. In this case I would recommend that this capacitor be changed to 30 or 40 pF and the output analyzed again.

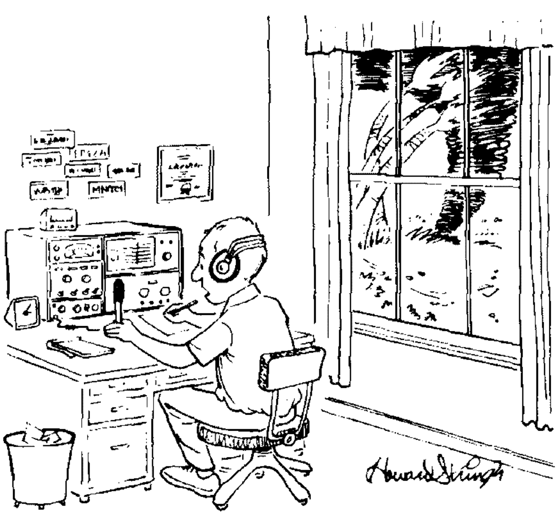
power supply

The regulated power supply in **fig. 2** has proven to be very satisfactory for operation of the Sonobaby and a solid-state whf-fm receiver. The parts are not at all critical – use whatever you have on hand. The transformer should be rated at about 1 ampere; the 13-volt zener diode can be taken from the audio board of the original Sonobouy transmitter. The transistor, Q1, should be rated at 5 or 10 watts. A dpdt relay with a coil resistance of about 100 ohms completes the system.

conclusion

The Sonobaby can be used as a miniature base station or it can be built into a walkie-talkie. There are many compact vhf-fm receivers on the market which are sold as police monitors; they can be coupled with the Sonobaby fm transmitter to provide a complete vhf-fm transmitter, and you can be on the air enjoying the benefits of 2-meter fm.

ham radio



"You say you're snowbound in a blizzard about twenty miles north of Notrees, Montana? Well, listen, I hate to interrupt, but . . ."

direct-reading capacitance meter for electrolytics

This simple adapter turns your vtvm or vom into a direct-reading capacitance meter for electrolytic capacitors up to 1000 uF

Herb Schoenbach, W9DJZ, 8 S. 275 Winwood Way, Apt. 6, Downers Grove, Illinois 60515

Some years ago I converted some transistor testers to give me direct readout of *beta* on the resistance scale of a vtvm or vom. This was a real time saver at home and on the job and provided accurate readings over two decades without any switching.[1] Since then I have been looking for a way to read voltages on the resistance scale. Imagine being able to read from 1 to 100 volts on the same scale with reasonable accuracy! Although I haven't found a way to do that it became apparent that capacitance could be read this way.

The required capacitance scale is the same as the standard resistance scale. My other capacitance meters (I've had two) cover values between 10 and 1000 μF and could not be modified to do so. The adapter I built for neasuring large values of capacitance is quite simple (see fig. 1). Fig. 2 shows a version for volt-ohmmeters which works well and is for those of you who don't own a vtvm.

construction

If you have a vtvm use the circuit of fig. 1. Volt-ohmmeters won't be as accurate since they load the circuit more and are not quite linear on ac. A 5000-ohm-per-volt ac meter is all right; a 1000-ohm-per-volt meter is usable but not accurate. In any case, the most critical part is the capacitor you use for a standard (C_S in the diagrams). Use a good quality 5% or better tantalum capacitor, even if you have to buy it.

The value of C_S must be equal to, or a decade multiple of, the exact midscale ohms reading. This corresponds to half-scale reading on any dc scale. The value you select will determine what center scale reading represents in microfarads. There's nothing wrong with making it a two-range unit if you wish. I used a 100 μF standard since my vtvm reads 10 at midscale; with this setup I can read capacitor values from 10 to 1000 μF very well.

The transformer can be a small filament type. Use as low a voltage (between 2 and 6 volts) as you can to match a 2-, 3-, or 5-volt ac range on your meter.

The resistance of the calibration pot should be selected experimentally for your particular meter. You should easily *be able to set full scale with it.* The calibrate switch is used only on the vtvm version since the vom version will read full scale whenever the test (C_X) terminals are open.

The remaining parts may or may not be necessary; let me explain their purpose and you can decide. C1 is a blocking capacitor. Most vtvms have one built in, in which case you don't need it. Voms might read correctly without it; one of mine does. C2 is an rf bypass; it is needed mostly on sensitive meters or if you are near a broadcast station or have your transmitter on.

The diode CR1 can be any small silicon diode or rectifier. It insures that a dc voltage will build up across the test

capacitor. It is not absolutely necessary because the normal rectifier action of electrolytic capacitors will do this to some extent. Readings are almost identical with or without the diode, but including it might make you feel better. At low voltages most electrolytics start to act like non-polarized devices; if you intentionally reverse polarity you get nearly the same reading.

I won't tell you how to put these parts in a box. There are enough good tips on construction practices in other articles — just go tò it.

operation

Here are some tips on using this gadget. Remember to set and connect your meter as an ac voltmeter on about a 3-volt range. *Do not* set to a resistance or ohmmeter position. But when you make your capacitance measurements read the capacitance value on the resistance (ohms) scale.

Check your full-scale calibration often if your line voltage fluctuates. If you use a 6-volt transformer tap do not test capacitors with less than a 10-volt rating. If you use a 3-volt tap you can safely test capacitors rated at 6 volts.

You will find that many electrolytics read almost double their marked value

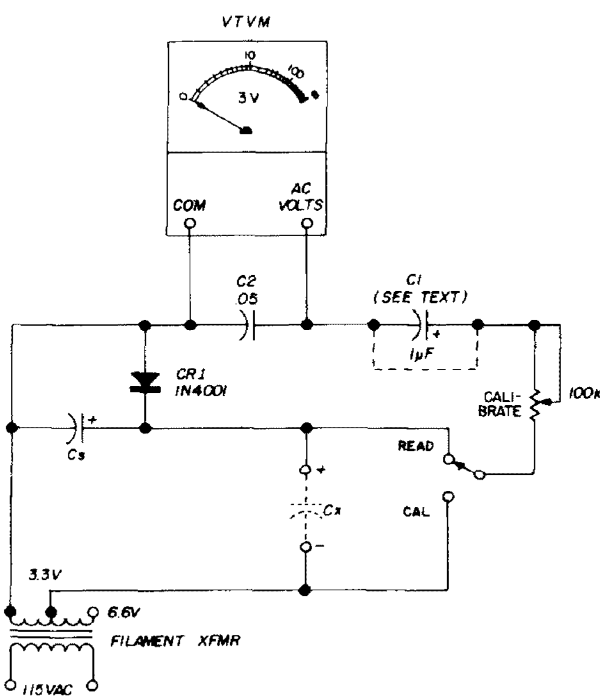


fig. 1. Capacitance meter adapter for a vtvm which has an ohms scale which increases upscale. To calibrate the adapter, push the calibrate switch and set the calibration pot for 3 Vac full-scale. Cs is a 5% tantalum capacitor, 10 volts or more (see text).

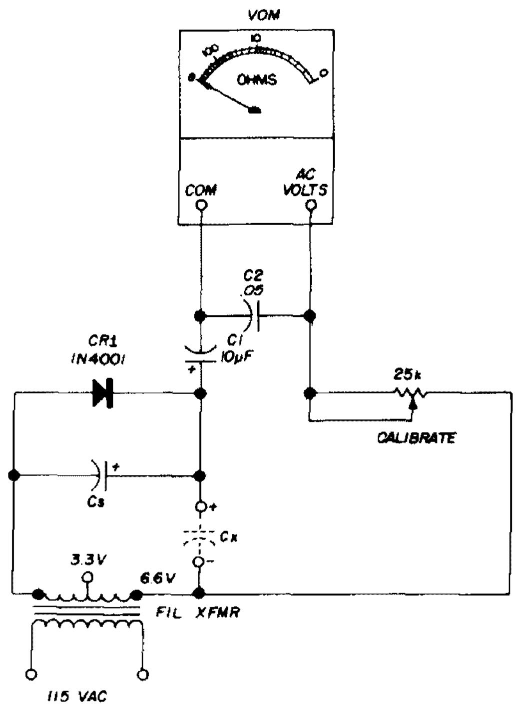


fig. 2. Capacitance meter adapter for volt-ohmmeter which has ohms increasing downscale. To calibrate the adapter set the calibrate pot for 5 Vac full-scale with capacitor Cx out of the circuit. Cs is a 5% tantalum capacitor, 10 working volts or more (see text).

when new. This is normal as capacitor tolerances are commonly plus 90%, minus 10%. High-voltage units may not read correctly unless they have been *formed* by operating them at their normal voltage for some time prior to testing.

Check unknown capacitors for shorts and leakage before trying to determine their capacitance. You will find leakage as low as 10,000 ohms does not affect accuracy much. You will want to check various known-value capacitors to be sure your adapter is working properly and to build up confidence in it. Even I was skeptical at first.

reference

1. Herb Schoenbach, W9DJZ, "Two More Transistor Testers," *73, September, 1969, page 133.*

ham radio

high-performance cw processor for communications receivers

Frequency modulating the telegraphy signals in your receiver provides an interesting and profitable addition to conventional receiver design

Dan E. Hildreth, W6NRW, 936 Azalea Drive, Sunnyvale, California 94086

There's no doubt that a properly designed cw receiver — of the current basic design — leaves little to be desired in terms of plucking out weak signals, providing the *feel* of the band, and allowing the character of the other man's transmitter to come through. But what about those times when impulse noise is heavy and you would like to have compression without distortion? The technique described here is something a brass-pounder can *add* to his present receiver to become a discriminating man! The system operates on received signals to give them the same character they would have if they had been tone frequency modulated. Detection is set up to demodulate an fm signal. In this way the advantages of an fm system are obtained without the necessity of transmitting fm.

When impulse noise is present the results of fm are well known. And when tuning across the band looking for small signals — or just general tuning — it's a relief to be able to tune across a strong signal without having it take off the top of your head. Finally, it is nice to have a choice, and the unit shown here makes it possible to quickly choose between three modes of operation:

1. Conventional bfo

2. Fm with bfo

3. Fm with fixed tone

To simplify the description a unique detection system will be shown; then the idea will be integrated into a conventional communications receiver.

basic theory

Consider a conventional receiver (bfo turned off) tuned to a cw signal blasting out a series of di-di-di-dahs. The signal can be heard going on and off, but there is no audio beat note until the bfo is turned on. Now, instead of turning on a bfo, suppose that a different method is used to provide the audio tone.

A block diagram of this different approach is shown in **fig. 1**. In this system the received signal is frequency modulated just like an rf carrier would be in a transmitter. **Fig. 2** expands the frequency modulator block to give you an idea of how fm is accomplished. This is the same basic technique used by Armstrong in the 1930s, and is often used to produce narrow band fm.

The incoming signal is split into two channels. One of the channels is fed through a double-sideband suppressed-

carrier device, phase shifted 90 degrees, and added to the other channel. The relative signal strength in each channel determines the degree of modulation.

The 90-degree phase shift could be tend to block small-signal inputs. To avoid that problem sufficient gain is used ahead of the multiplier to get the noise background up to at least 10 millivolts at the input to the multiplier. A fullwave

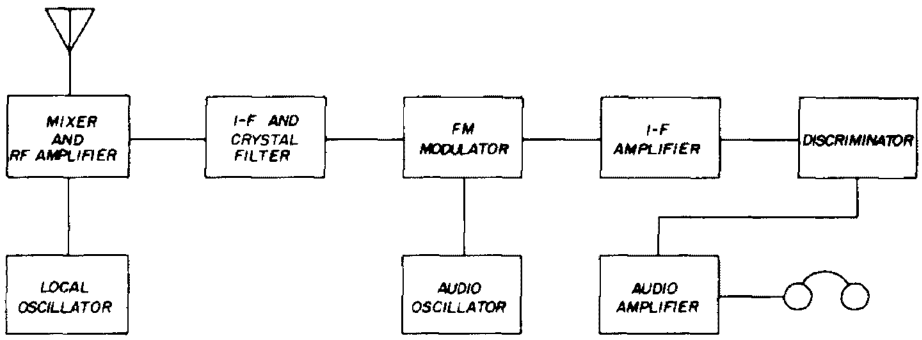


fig. 1. Basic system for frequency modulating telegraphy signals in a communications receiver.

placed in either channel – but not both. If no phase shift were present the system would simply supply amplitude modulation. In either case the detected tone is identical with what is fed into the modulating port of the double-sideband suppressed-carrier circuit.

the circuit

To a radio amateur the sound of a receiver like fig. 1 is almost as sterile as a code-practice oscillator, so let's bear with a little more complication to get the *feel* back. The arrangement shown in **fig. 3** provides the three modes of operation and may be easily added to an existing receiver.

When the modulator is excited by the receiver's normal bfo, the *feel* is back! Then, for maximum quiet and/or very weak signals, the modulator can be switched to the input to the audio oscillator input. **Fig. 4** shows a schematic of the complete system.

At this point you may note that a considerable amount of energy is left in the carrier with sideband information some 6 dB down. To reduce that problem a final system should multiply the frequency of the signal before running it through the discriminator.

frequency multiplier

Conventional frequency multipliers rectifier circuit with forward bias on the diodes reduces diode dead space to enable frequency multiplication to start with very small signals. (If you have worked with nbfm you are well aware that frequency multiplication will increase the modulation index, thus improving threshold signal detection.) Although more multiplication than is shown here would be beneficial for an optimum system you could obtain an improvement of only 2 or 3 dB at best.

discriminator

I chose a Travis discriminator circuit because it does not require a special transformer – a problem in this case because of the odd frequency. (A Travis discriminator could be thought of as a "push-pull staggered" slope detector.) With careful construction and adjustment of the balance potentiometer a-m rejection is reasonably good.

The easiest discriminator tuning procedure is as follows (refer to fig. **4**): Energize the system with a strong, steady carrier at the center i-f frequency. Peak L1 while looking at the rectified dc voltage at diode CR1. Note the peak voltage; then tune L1 to raise the circuit frequency until the voltage is one-half of peak. Repeat the same procedure (opposite polarity) with L2 while looking at the voltage at diode CR2 but offset this

resonant circuit to the low side of peak voltage.

Assuming an i-f selectivity such as that recommended under *receiver requirements* is used the discriminator will be primarily to hold shielding requirements to a reasonable level. It should be pointed out that the second harmonic of the i-f is in the broadcast band, so care should be taken to prevent pick-up from the local

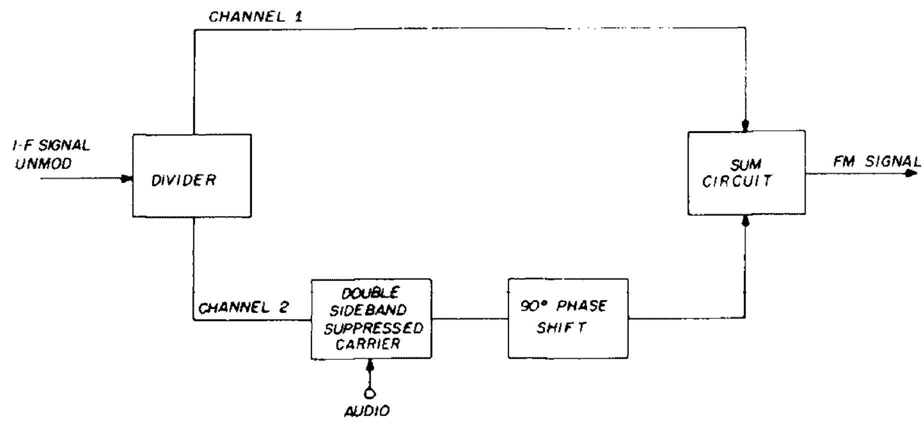


fig. 2. Basic phase modulation system. The rf input signals to the summing circuit should be approximately equal when full audio modulation is present.

relatively wide. This avoids the spurious detection on each side of a discriminator's S-curve that is sometimes annoying in tuning fm signals.

An fet was used at the input to avoid discriminator loading. This stage is followed by just enough transistor gain to get reasonable headset volume. The circuitry shown in **fig. 4** will easily drive an audio power stage if speaker operation is desired.

All of the circuits used in this system are rather conservative, with minimum use of LC elements. This was done

broadcast station. Normal point-to-point wiring is completely adequate for this circuit.

receiver requirements

While this system will function when used with a receiver without good i-f skirt selectivity the performance would be disappointing. To work well this system should be coupled to a receiver with a good crystal or mechanical filter. Present indications are that an i-f bandwidth of 200 to 400 Hz, with skirts only twice that wide 60-dB down, will give excellent

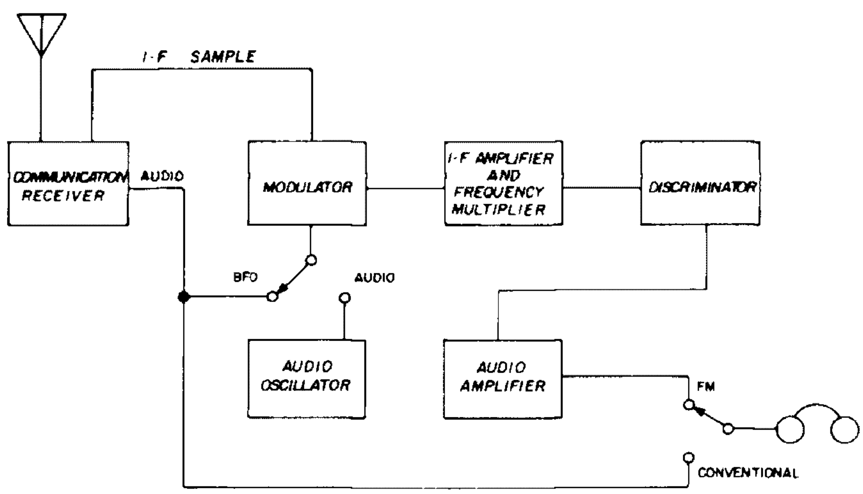


fig. 3. Signal processor is easily added to an existing receiver to provide new as well as conventional modes of cw reception.

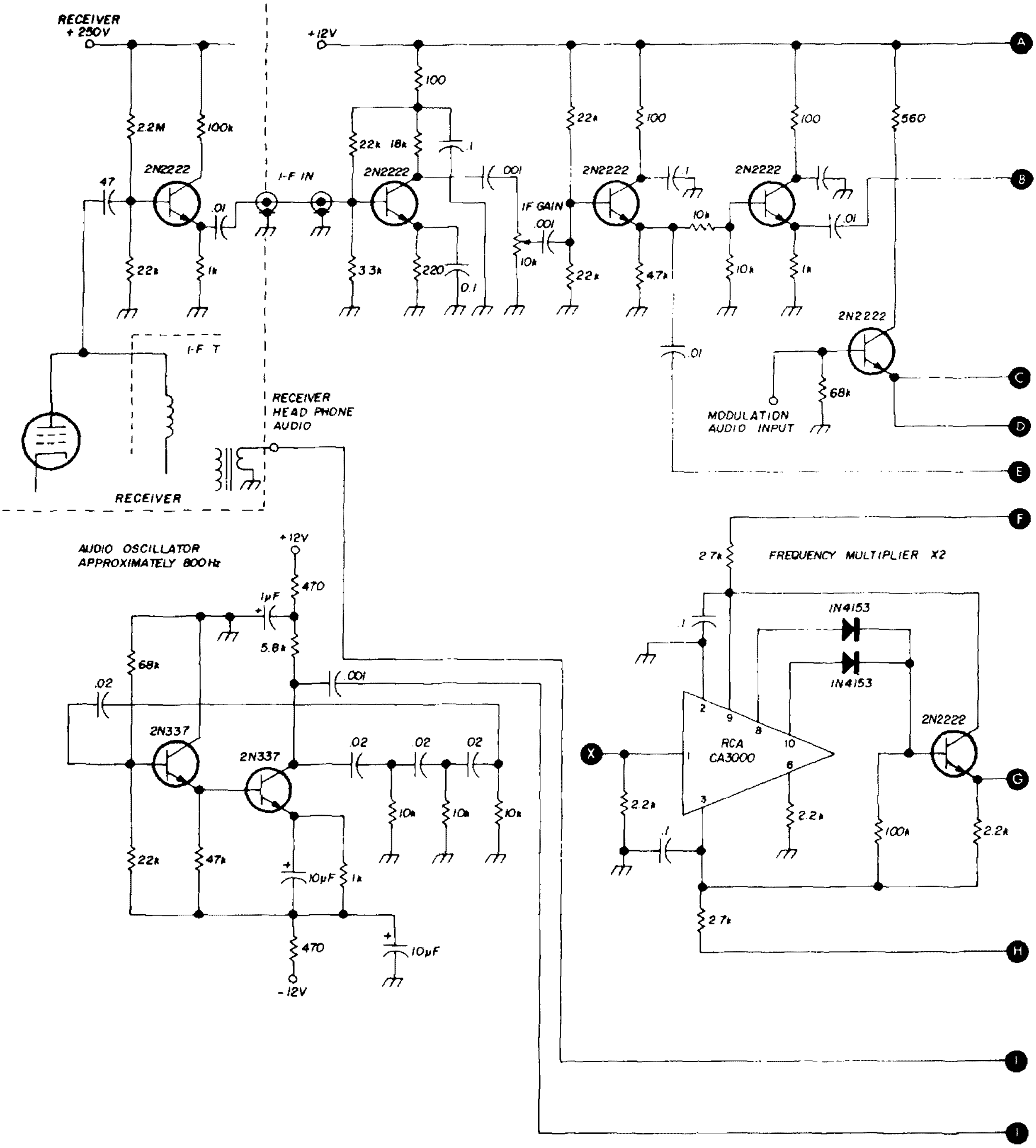


fig. 4. Schematic diagram of the cw signal processor.

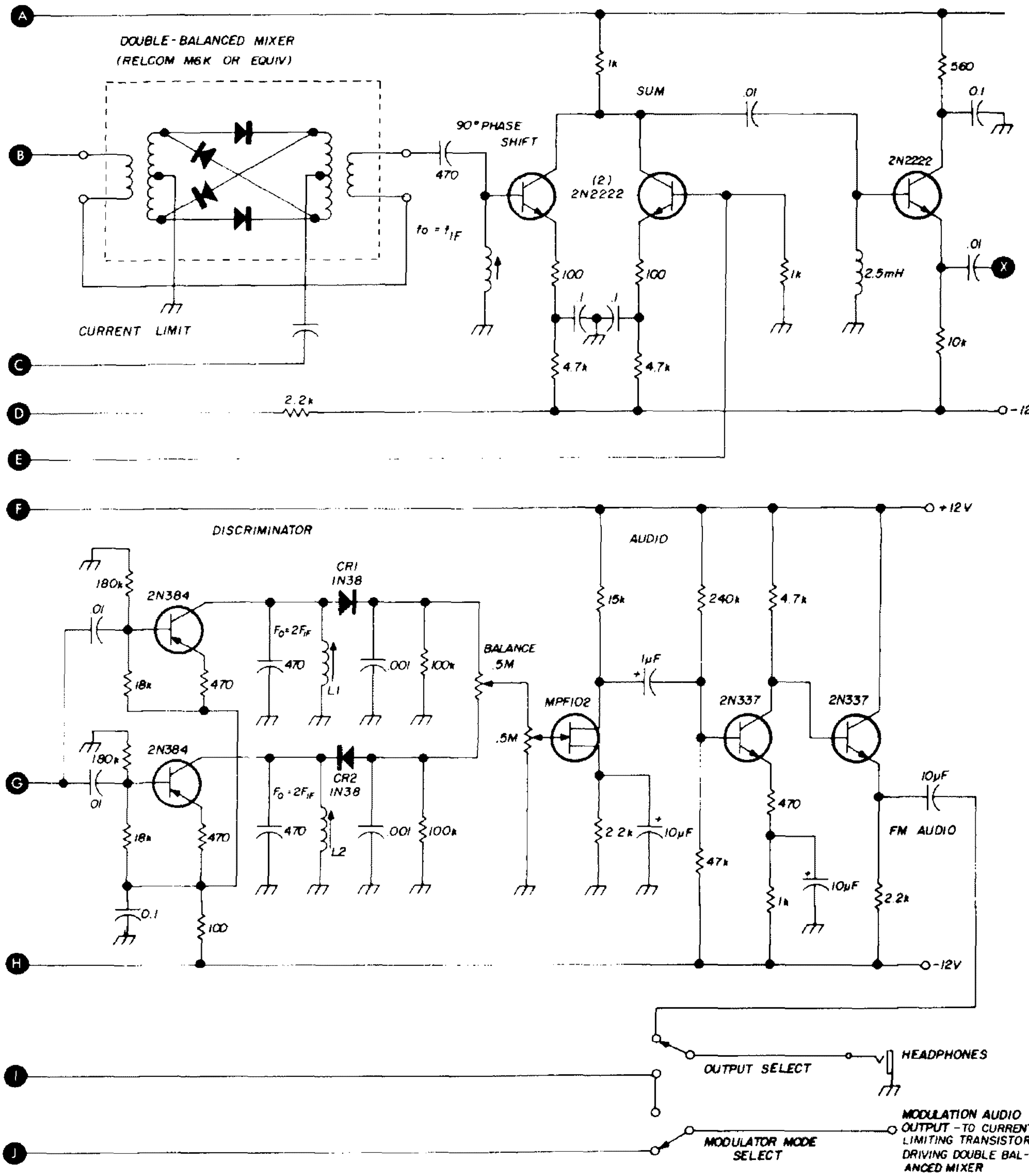
results.

It is important to note that the coupling point should be right after the crystal filter (which is usually right after the mixer). This prevents the bfo oscillator signal from getting into the input of the frequency modulator.

summary

Aside from the improved results obtained with this system in the presence of impulse noise, it is quite a treat to my tired old ears to be able to tune across the band, hearing weak signals and at the same time, not having my ear drums drilled out when that kilowatt down the street comes on.

Although the system shown in fig. 4 will not outperform a conventional bfo/product detector when low background



noise is present there is hope with future extensions of the basic idea.

The next step, of course, is to replace the discriminator with a phase-locked loop. Once this is done, a narrow-band audio filter may be used with more advantage than is gained when filters are placed after a conventional a-m or fm detector.

This frequency-modulated cw reception has proven interesting to me, but it is relatively complex, and the system with the phase-locked loop will be even more so. (I am aware that there are easier ways to phase modulate a signal but I want two separate channels — I have plans there too!)

This article has stopped with the basic idea. With enough interest there will be more to come. The phone operators have had interesting new modulation methods to play with; isn't it about time the brass pounders had a new toy?

ham radio

the application of stress analysis to antenna systems

Understanding the mechanical design of antennas and supporting structures

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Most amateurs at one time or another have wished to extend a mast a few feet higher or make a beam with longer elements. The question is, "How high or how long can one safely go?" The calculations involved in arriving at the answer to this question are relatively simple, and a good feeling for the problem may be had with little "stress or strain" on the part of the interested ham.

basic data

The problem simply involves comparison and matching of material strength with maximum anticipated load. Two equations are used to develop the comparison:

$$M_L = F \times D \text{ and } M_R = f \times S$$

where M_L = bending moment developed by loading force (lb)

M_R = restraining moment developed by loaded structure (lb)

F = force of the load (lb)

D = distance from the effective point of application of F to the fulcrum (in.)

f = strength of the material being used ($lb/in.^2$)

I = moment of inertia of the cross section about its neutral axis ($in.^4$)

C = distance from the neutral axis to the extreme fiber (in.)

S = section modulus ($in.^3$) = I/C

Some of these terms may not be too familiar and their calculation even less so. However, the table and chart makers of America have provided numerous graphic aids that do almost everything short of coming up with the final answer.

bending moment

Since the problem revolves about the "bending moment" concept, let's consider first what bending moment means and how it's computed. A bending moment is the product of a force and its distance to the point under consideration, where the force is the component perpendicular to a line drawn from the point of application to the point under consideration.

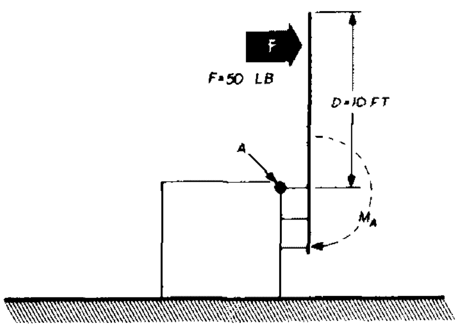

fig. 1. The bending-moment concept. Force F is a wind load exerted on a mast extending 10 feet above its last support.

example 1: In fig. 1 the bending moment at point A is:

$$M_A = F \times D$$
$$= 50 \text{ lb} \times 10 \text{ ft.}$$
$$= 500 \text{ ft-lb.}$$

This might represent the bending moment caused by a small antenna, with a 50-lb wind load, mounted on a mast that extends 10 feet above its last support, or the tension in a wire antenna supported by the mast.

The question now arises, "Fine, but where did the wind-load value come from?" This again is relatively easy to calculate; or if you're lucky, the job may have been done already by the manufacturer of your antenna or tower. The calculation goes like this:

$$F_{WL} = A_e \times P_W$$

where F_{WL} = wind load force

A_e = effective area on which the wind is acting

P_W = wind pressure

table 1. Summary of parameters used in bending-moment analysis examples.

wind pressure

wind velocity (mph)	pressure on flat surface (lb/ft^2)*
10	.42
20	1.7
30	3.8
40	6.7
50	10.5
60	15.1
70	20.6
80	26.8
90	34.0
100	42.0

area factors

type of cross section		multiplication factor
open face (latticed) — square —	on face	2.20
	on corner	2.40
triangle —	on face	2.00
	parallel to face	1.50
closed face (solid) —	square or rectangular	1.00
	hexagonal or octagonal	0.80
	round or elliptical	0.60

*In most instances 20 lb/ft^2 is considered a safe value

Further explanation of A_e and P_W are as follows. The effective area is that of a flat plate which, if substituted for the object of interest, would accumulate the same total force. It is recommended that anyone working on this problem consult a handbook such as Reference 1 to obtain multiplication factors for different cross sections and wind-load values for different wind velocities. These values are summarized in table 1, and an example of their use follows.

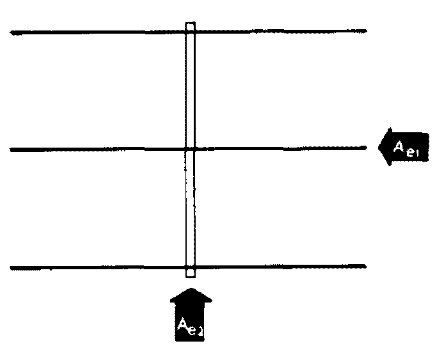

fig. 2. Typical beam antenna used in the wind-loading example. Boom is 2 in. dia. x 26 ft; elements are 1 in. dia. x 36 ft.

example 2 Find the wind load on a 3-element 20-meter beam (see **fig. 2**).

$$A_{e2} = 2 \text{ in.} \times \frac{1 \text{ ft.}}{12 \text{ in.}} \times 26 \text{ ft} \times .6 = 2.6 \text{ ft}^2$$

$$A_{e1} = 3 \times 1 \text{ in.} \times \frac{1 \text{ ft}}{12 \text{ in.}} \times 36 \text{ ft} \times .6 = 5.4 \text{ ft}^2$$

$$A_{e_{max}} = \sqrt{A_{e1}{}^2 + A_{e2}{}^2} = \sqrt{36} = 6 \text{ ft}^2$$

Using 20 lb/ft^2 wind load,

$$F_{WL} = 6 \text{ ft}^2 \times 20 \text{ lb/ft}^2 = 120 \text{ lb}$$

restraining moment

Now we'll examine the restraining *moment*, which is the force with which the support structure resists being bent by the loading forces. The equation for this moment, as given above, is:

$$M_R = \frac{fxI}{C} = fxS$$

The determination of the values to use in this equation has again been made easier by charts and tables available in many structural and mechanical handbooks. Some of the most common values are shown in **tables 2** and **3** and **fig. 3.**

table 2. Strength of materials.

material	tensile strength (lb/in.^{2})	compressive strength elastic limit (lb/in.^{2})
wrought iron	30,000	30,000
structural steel	40,000	40,000
copper (hard)	42,000	——
brass (hard)	60,000	——
aluminum-extruded tubing less than ¼" wall;		
2024-T4	40,000	36,000
6061-T6	35,000	35,000
6063-T6	25,000	25,000
7178-T6	76,000	76,000
white oak	——	5,000
yellow pine	——	4,000

example 3 Now we can take a look at the problem of extending our mast above the tower to get the psychological advantage of a few extra feet of height. Assume the three-element 20-meter beam of example

table 3. Elements of section.

cross section		lb/ft	I	S	C
½ diameter	.031 wall	0.054	0.001	0.005	0.25
½	.062	0.101	0.002	0.008	0.25
½	.125	0.173	0.003	0.012	0.25
¾	.062	0.159	0.008	0.021	0.375
¾	.125	0.288	0.012	0.033	0.375
1	.062	0.216	0.020	0.041	0.500
1	.125	0.405	0.034	0.067	0.500
1¼	.062	0.274	0.041	0.071	0.625
1¼	.125	0.520	0.071	0.113	0.625
1½	.062	0.332	0.073	0.097	0.750
1½	.125	0.635	0.129	0.172	0.750
2	.062	0.447	0.179	0.179	1.000
2	.125	0.865	0.325	0.325	1.000
2	.250	1.616	0.537	0.537	1.000
3	.125	1.328	1.169	0.779	1.500
3	.250	2.540	2.059	1.373	1.500
4	.125	1.790	2.859	1.429	2.000
4	.250	3.463	5.200	2.600	2.000

I = moment of inertia (in.^{4})
S = section modulus (in.^{3})
C = distance from the neutral axis to the extreme fiber (in.)

Reference: *Alcoa Structural Handbook,* Aluminum Company of America, Pittsburg, Pennsylvania.

2 and assume a heavy-duty mast of 2-in. O. D. X. 1/4-in. wall steel tubing. From fig. 4,

F_a = antenna wind load

F_m = mast wind load

D = height of mast

d = mast diameter

The effect of mast wind loading is evenly distributed along the mast; therefore this force may be considered to act at a point half-way up the mast.

Maximum allowable bending moment:

$$M = f \times S$$

$$f = 40{,}000 \frac{lb}{in.^2} \text{ (from table 2)}$$

$$S = 0.537 \text{ in.}^3 \text{ (from table 3)}$$

$$= 40{,}000 \frac{lb}{in.^2} \times 0.537 \text{ in.}^3$$

$$= 21{,}500 \text{ in.-lb} \times \frac{ft}{12 \text{ in.}} = 1{,}800 \text{ ft-lb.}$$

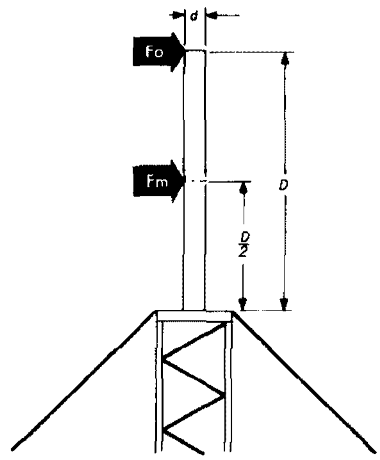

fig. 4. Extending a mast above the tower. Parameters are defined in the text.

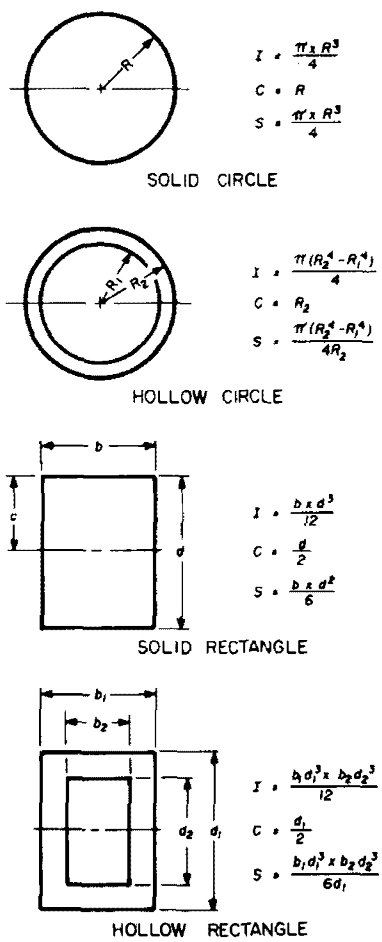

fig. 3. Data for determining I, S, and C of different shapes.

Antenna wind load:

$$F_a = 6 \text{ ft}^2 \times 20 \frac{lb}{ft^2}$$

$$= 120 \text{ lb}$$

Mast wind load:

$$F_m = D \times d \times .6 \times 20 \frac{lb}{ft^2}$$

$$= D \times .167 \times .6 \times 20$$

$$= 2D$$

Total bending moment:

$$M_{total} = F_a \times D + F_m \times \frac{D}{2}$$

$$= 120 \text{ lb} \times D + 2D \times \frac{D}{2}$$

$$= 120D + D^2 \text{ (ft.-lb.)}$$

Maximum allowable mast height:

Setting the maximum allowable bending moment equal to the total bending moment, we can now solve for the maximum allowable mast height.

$$1{,}800 = 120D + D^2$$

$$D^2 + 120D - 1{,}800 = 0$$

Using the general solution for a quadratic equation,

$$\frac{-b \pm \sqrt{b^2 - 4ac}}{2a}$$

supported at the center and designed to withstand 70-mph wind loading. Since the element is supported at the middle, we need consider only a half element (fig. 5).

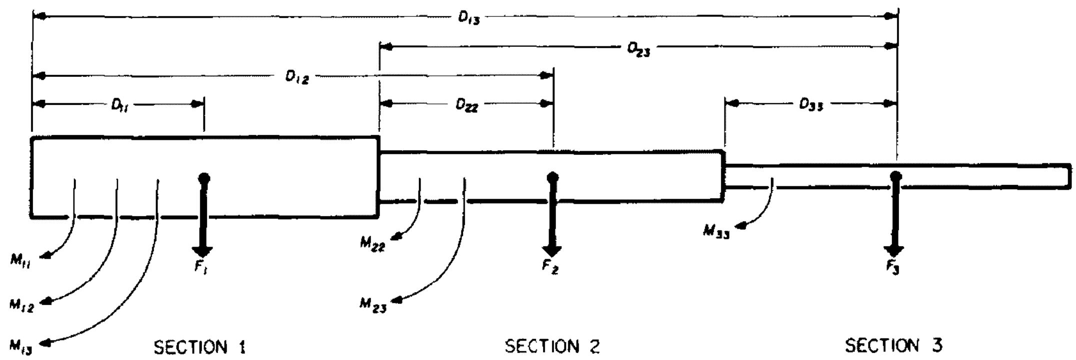

fig. 5. Parameters for a typical beam element designed to withstand 70-mph wind loads.

where $a = 1$

$b = 120$

$c = -1{,}800$

we have

$$D = \frac{-120 \pm \sqrt{120^2 - 4 \times 1 \times (-1{,}800)}}{2}$$

$$= \frac{-120 \pm 147}{2} = \frac{27}{2}$$

$= 13.5$ ft maximum

A safety factor to suit the taste of the designer may now be applied to come up with the final height. If it is desired to allow for contingencies such as ice loading or inferior materials, the more conservative approach would be to use values in the initial calculations to account for them and not try to lump them all into a magic safety factor.

Use the safety factor to account for contingencies you have not thought of yet! If such an idea should appeal to you, be sure to consider also the allowable bending moment for your tower (should be available from manufacturer) and the added load on your rotator (15 ft of 2-in. diameter 1/4-in. wall steel tubing will weigh about 70 lb).

example 4 Next let's look at the design of a beam element, say a 20-meter element

Section 3: ½-in. diameter, 0.031-in. wall

$$M = S \times f$$

$$= .005 \text{ in.}^3 \times 35{,}000 \frac{\text{lb.}}{\text{in.}^2}$$

$$= 175 \text{ in.-lb} \times \frac{\text{Ft}}{12 \text{ in.}}$$

$= 14.6$ ft-lb allowable

wind loading per linear foot

$$= .5 \text{ in.} \times \frac{\text{ft}}{12 \text{ in.}} \times .6 \times 20 \frac{\text{lb}}{\text{ft}^2}$$

$$= .5 \frac{\text{lb}}{\text{ft}}$$

weight per linear foot

$$= .054 \frac{\text{lb}}{\text{ft}}$$

total load per linear foot

$$= .554 \frac{\text{lb}}{\text{ft}}$$

If section 3 is 7 feet long:

$$F_3 = 7 \text{ ft} \times .554 \frac{\text{lb}}{\text{ft}}$$

$= 3.87$ lb

$$M_{33} = D_{33} \times F_3$$

$= 3.5$ ft x 3.87 lb

$= 13.5$ ft-lb loaded

This is a safe loading when compared to the calculated allowable load of 14.6 ft-lb.

Section 2: ¾-in. diameter, 0.062-in. wall

$$M = S \times f$$

$$= .021 \text{ in.}^3 \times 35{,}000 \frac{\text{lb}}{\text{in.}^2}$$

$$= 735 \text{ in.-lb} \times \frac{\text{ft}}{12 \text{ in.}}$$

Total moment at 2-3 = 61.2 ft-lb allowable

wind loading per linear foot

$$= .75 \text{ in.} \times \frac{\text{ft}}{12 \text{ in.}} \times .6 \times 20 \frac{\text{lb}}{\text{ft}^2}$$

$$= .75 \frac{\text{lb}}{\text{ft}}$$

weight per linear foot

$$= .159 \frac{\text{lb}}{\text{ft}}$$

total load per linear foot

$$= .909 \frac{\text{lb}}{\text{ft}}$$

If section 2 is 6 feet long:

$$F_2 = 6 \text{ ft} \times .909 \frac{\text{lb}}{\text{ft}}$$

$$= 5.45 \text{ lb}$$

$$M_{22} = F_2 \times D_{22}$$

$$= 5.45 \text{ lb} \times 3.0 \text{ ft} = 16.4 \text{ ft-lb}$$

$$M_{23} = F_3 \times D_{23}$$

$$= 3.87 \text{ lb} \times (6 + 3.5) \text{ ft} - 36.8 \text{ ft-lb}$$

Total moment at 1-2 = 53.2 ft-lb

This is a safe loading when compared to the calculated allowable load of 61.2 ft-lb.

Section 1: 1-in. diameter, 0.062-in. wall

$$M = S \times f$$

$$= .041 \text{ in.}^3 \times 35{,}000 \frac{\text{lb}}{\text{in.}^2}$$

$$= 1440 \text{ in.-lb} \times \frac{\text{ft}}{12 \text{ in.}}$$

$$= 120 \text{ ft-lb allowable}$$

wind loading per linear foot

$$= 1 \text{ in.} \times \frac{\text{ft}}{12 \text{ in.}} \times .6 \times 20 \frac{\text{lb}}{\text{ft}^2}$$

$$= 1.0 \frac{\text{lb}}{\text{ft}}$$

weight per linear foot $= .216 \frac{\text{lb}}{\text{ft}}$

total load per linear foot $= 1.216 \frac{\text{lb}}{\text{ft}}$

If section 1 is 5 feet long:

$$F_1 = 1.216 \frac{\text{lb}}{\text{ft}} \times 5 \text{ ft}$$

$$= 6.08 \text{ lb}$$

$$M_{11} = F_1 \times D_{11}$$

$$= 6.08 \text{ lb} \times 2.5 \text{ ft} = 15.2 \text{ ft-lb}$$

$$M_{12} = F_2 \times D_{12}$$

$$= 5.45 \text{ lb} \times (5 + 3.0) \text{ ft} = 43.6 \text{ ft-lb}$$

$$M_{13} = F_3 \times D_{13}$$

$$= 3.87 \text{ lb} \times (5 + 6 + 3.5) \text{ ft} = 56.0 \text{ ft-lb}$$

Total moment at center of element = 114.8 ft-lb

This is a safe loading when compared to the calculated allowable load of 120 ft-lb.

From these calculations the general method may be seen for solving this type of problem. It should be noted that the size of tubing used for each section was arrived at by trial and error; i. e., picking a size, calculating the allowable moment and loaded moment for a reasonable length, until a good and reasonable combination was obtained.

If it's desired to account for possible ice loading, simply increase the element diameter for the wind loading calculation and compute the weight of the ice and add it to the weight of the tubing (Weight of ice $= 62.5 \frac{\text{lb}}{\text{ft}^3} \times .9 = 56.2 \frac{\text{lb}}{\text{ft}^3}$).

If you wish to design a self-supporting vertical antenna, the same technique is also used. In this case, however, the weight causes compressive stress rather than bending moments and in most cases may be neglected.

example 5 As another example, therefore, let us look at the design of a 20-meter self-supporting vertical, 18 feet high, allowing for 70 mph wind *and* 1/4 in. of radial ice. **Fig. 6** illustrates the parameters.

Section 3: ½-in. diameter, 0.125-in. wall, (plus ¼ in. of ice)

$$M = S \times f$$

$$= .012 \text{ in.}^3 \times 35{,}000 \frac{\text{lb}}{\text{in.}^2}$$

$$= 420 \text{ in.-lb} \times \frac{\text{ft}}{12 \text{ in.}}$$

$$= 35 \text{ ft-lb allowable}$$

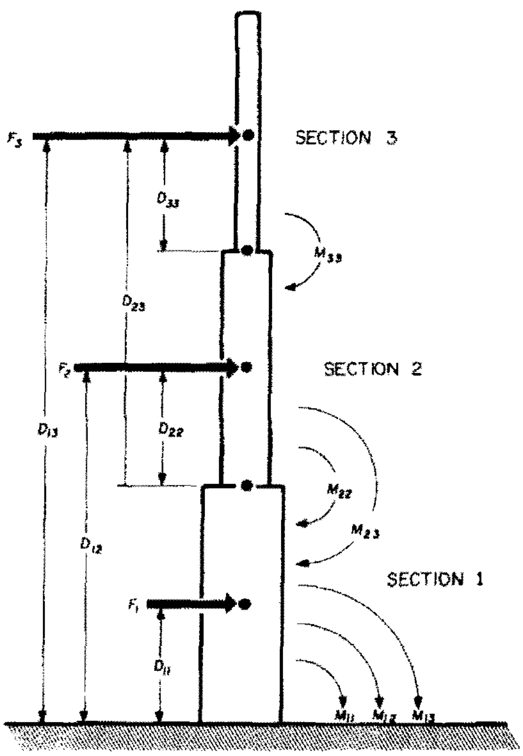

fig. 6. Parameters for a self-supporting vertical antenna allowing for 70-mph wind and ¼ in. of ice buildup.

wind loading per linear foot

$$= 1 \text{ in.} \times \frac{\text{ft}}{12 \text{ in.}} \times .6 \times 20 \frac{\text{lb}}{\text{ft}^2}$$

$$= 1.0 \frac{\text{lb}}{\text{ft}}$$

If section 3 is 8 feet long:

$$F_3 = 1 \frac{\text{lb}}{\text{ft}} \times 8 \text{ ft}$$
$$= 8 \text{ lb}$$

$$M_{33} = D_{33} \times F_3$$
$$= 4 \text{ ft} \times 8 \text{ lb}$$
$$= 32 \text{ ft-lb}$$

This is safe loading when compared to the calculated allowable load of 35 ft-lb.

Given the procedure for finding the wind load with 1/4 inch of ice on section 3 of the example shown in **fig. 6**, I'll leave as an exercise the calculation of moments for the other two elements. Assume the following: section 2: 3/4-in. diameter. 0.125-in. wall, 5 ft long section 1: 1-in. diameter, 0.125-in. wall, 5 feet long

If your calculations are correct, you should obtain a value of 177.5 ft-lb at the base, which is a safe load compared to the calculated allowable of 195 ft-lb.

compressive stress

Assume 1-in. diameter elements with 1/4-in. ice load and 18 feet high:

$$C = \frac{W}{A}$$

where C = compressive stress

W = total weight

A = cross section area of restraining material

weight of tubing

$$= 0.405 \frac{\text{lb}}{\text{ft}} \times 18 \text{ ft}$$
$$= 7.3 \text{ lb}$$

weight of ice

$$= \pi({r_2}^2 - {r_1}^2)\text{in.}^2 \times \frac{\text{ft}^2}{144 \text{ in.}^2} \times 18 \text{ ft.} \times 56.2 \frac{\text{lb}}{\text{ft}^3}$$

$$= \frac{3.14(.75^2 - .50^2) \times 18 \times 56.2}{144}$$

$$= 6.9 \text{ lb}$$

total weight $= 7.3 \text{ lb} + 6.9 \text{ lb}$

$$= 14.2 \text{ lb}$$

$$A = \pi({r_2}^2 - {r_1}^2)\text{in.}^2$$
$$= 3.14(.500^2 - .375^2)\text{in.}^2$$
$$= .344 \text{ in.}^2$$

$$C = \frac{W}{A}$$

$$= \frac{14.2 \text{ lb}}{.344 \text{ in.}^2}$$

$$= 41.3 \frac{\text{lb}}{\text{in.}^2}$$

This is negligible when compared to the strength of aluminum at $35{,}000 \frac{lb.}{in.^2}$

From this example we observe two important characteristics. First, the additional wind area caused by ice loading results in the requirement for heavier construction than the beam element of example 4 despite its being supported vertically instead of horizontally. Second, the compressive stress in the vertical element, even with ice loading, is quite negligible.

If you would like to design a husky beam element, you might try designing for 70 mph wind with simultaneous 1/4 in. of ice!

guying

Another problem for many hams is determining the proper size cables to use for guying an antenna support structure. A simplified approach will be taken to this problem, which should allow adequate safety.

To deal with this problem it's necessary to understand the concept of resolving a force into components. The reader is urged to refer to almost any high school physics book to review this technique.

Let's now consider the problem of guying a 65-foot tower that's supporting the antenna in example 2. See **fig. 7.**

example 6 Force of wind on tower:

area of one face = 2(1 in. x 12 in.) = 24 in.^{2}

¼ in. (12 in. + 16 in.) = 7 in.^{2}

total = 31 in.^{2}

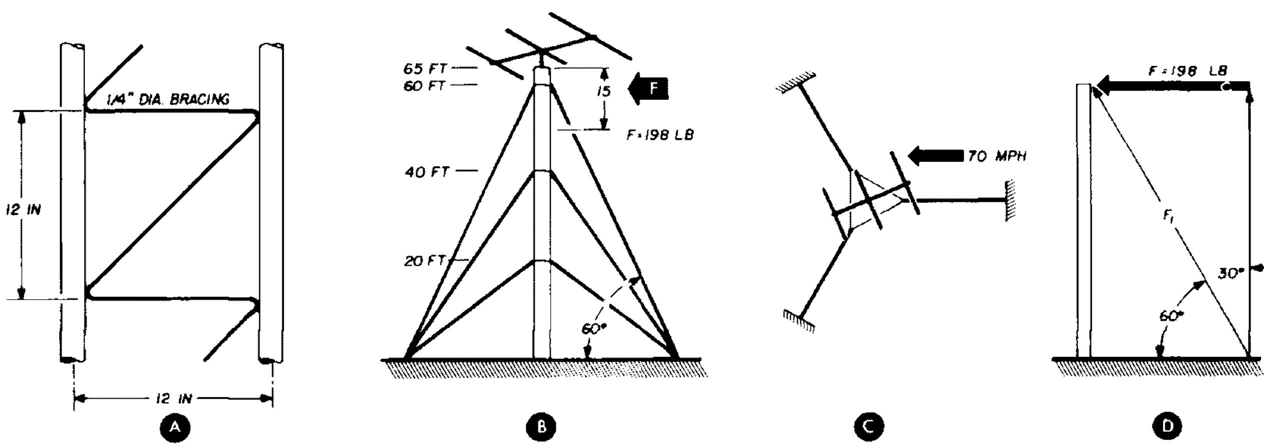

fig. 7. The guying problem for a 65-ft. tower supporting a typical 3-element beam. In A, tower elements are 1 in. dia. with ¼ in. bracing. B and C illustrate text description of wind load on top guys. D is geometry for calculating force on the top guy.

effective area of one face (cylindrical surfaces)

$= .6 \times 31 \text{ in.}^2$

$= 18.6 \text{ in.}^2$

effective wind area of open latticed triangular tower with wind applied perpendicular to a face = 18.6 in.^{2} x 2

$= 37.2 \text{ in.}^2 \frac{\text{ft.}}{144 \text{ fn.}^2}$

effective total wind area per linear foot of tower = $.258 \frac{ft^2}{ft}$

Tower wind load per linear foot

$= .258 \frac{ft^2}{ft} \times 20 \frac{lb}{ft^2}$

$= 5.16 \frac{lb}{ft}$

The top set of guys will have to take all of the wind force above that set of guys plus the antenna, and about half the force between the top and middle set of guys (**fig. 7B**).

wind load of antenna (from example 2)
$= 120$ lb

wind load of tower $= 5.16 \frac{\text{lb}}{\text{ft}} \times 15$ ft
$= 78$ lb

total wind force $= 198$ lb

Component of force in the top front guy (see fig. 7C):

$$F_1 = \frac{198 \text{ lb}}{\sin 30^\circ} = \frac{198 \text{ lb}}{.50} = 396 \text{ lb}$$

Added to this would be about 100 lb of force from tightening up the guy, or a total of just about 500 lb of tension. Therefore, using 1/8 in. aircraft cable with a breaking strength of 2100 lb would offer a comfortable safety margin. The strain on the lower guys would be somewhat less: the calculation is left as an exercise for the reader.

When selecting guying cable, careful attention should be paid to the load rating. There are many different types of 1/8 in. cable — some flexible stainless steel, some flexible winching cables with fiber core, less-flexible 5- or 7-strand galvanized, and of course the single strand type — all with different load ratings. Also, a given cable will have many ratings depending on the service; i. e., breaking strength, yield strength, and working strength. And for that matter, the working strength will vary depending on the intended use; e. g., winch service rating would be lower than that for guying service. Suffice to say — check with your supplier and be sure of what you're buying.

These are the basics. If you've been guessing up to now, you might try some calculations to see how safe you are. If you've been putting off a project because of fear of disaster, perhaps you can try it now (with a *little* less fear)!

reference

1. *Reference Data for Radio Engineers,* International Telephone and Telegraph Corp., New York, New York.

ham radio

solid-state RTTY monitor scope

This solid-state RTTY monitor scope was designed specifically as a companion to the ST-6 terminal unit

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In the past amateurs who turned to RTTY usually used a vacuum-tube type terminal unit such as the TTL/2 or the circuit designed by W2PAT. Now the trend is toward the high performance solid-state ST-5 or ST-6 designed by W6FFC.[1,2] When I decided to go solid state I decided to go all the way and include a monitor scope that was based on solid-state components.

Although the oscilloscope circuit shown in **fig. 1** is designed primarily for use with the ST-5 or ST-6 it can be used with any RTTY terminal unit as long as the vertical and horizontal amplifiers are not over-driven.

circuit

The amplifier circuits are quite simple. In the vertical amplifier an emitter-follower input stage (Q1) provides high input impedance. The CRT driver stage (Q3) is rated at 250 volts. Vertical gain is controlled by the 5000-ohm pot (R9) in the emitter circuit of the driver stage.

The horizontal amplifier circuit is identical to the vertical lineup. Note in **fig. 1** that even-numbered components are part of the horizontal amplifier chain; components in the vertical amplifier are designated by odd numbers. The gain of

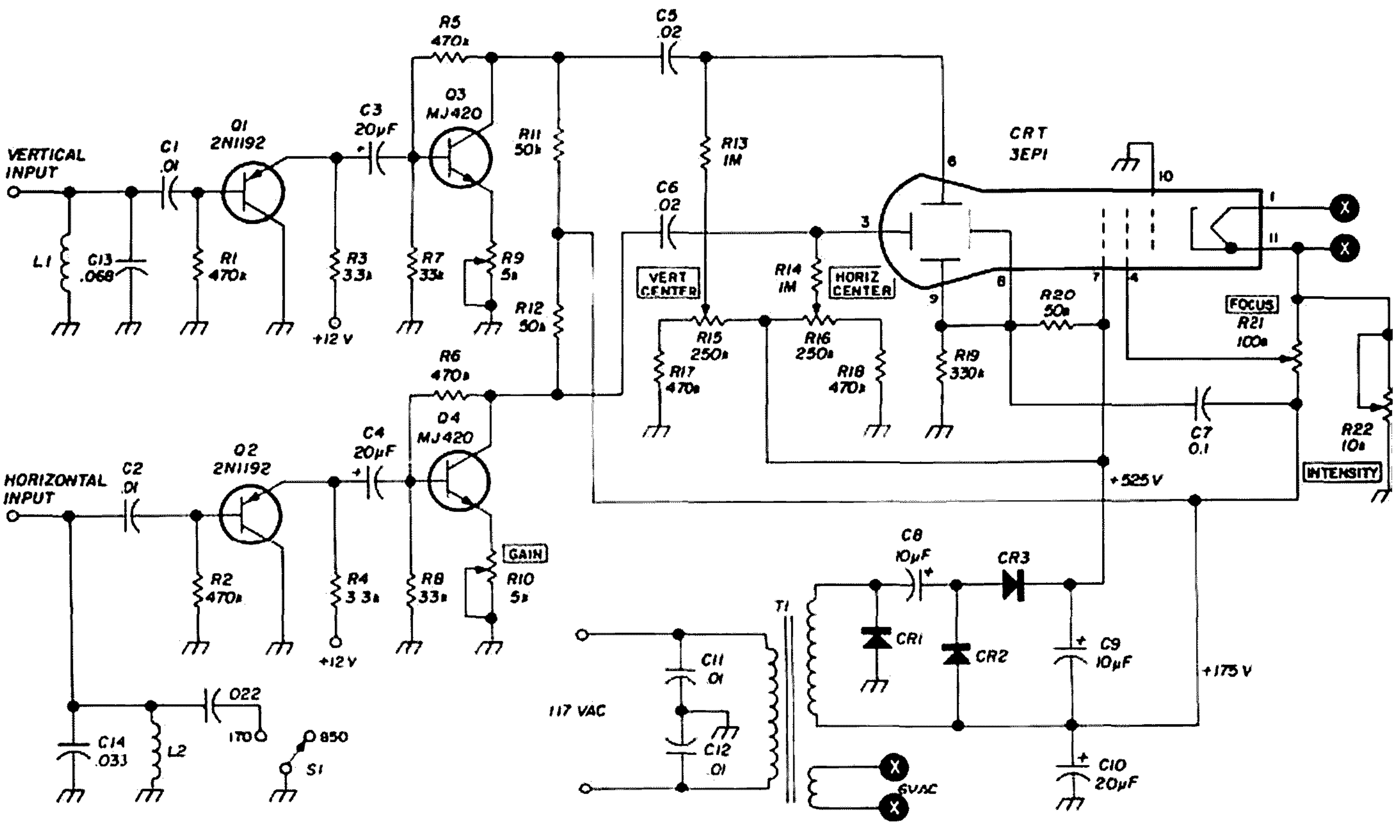

fig. 1. Solid-state RTTY monitor scope is designed for the ST-6 RTTY terminal unit.

the amplifier system is more than adequate to fill the screen of a 3-inch CRT.

The filters used in the Mainline ST-5 and ST-6 terminal units result in a very poor oscilloscope display since there is voltage on the *mark* toroid when a *space* signal is present, and vice versa. When the ST-6 is used on 850 shift the scope display looks like two bananas; on 170 shift the display looks like two fat footballs.

For example, when tuned to *mark* there is still vertical deflection voltage which opens up the horizontal trace. The higher the vertical voltage, of course, the more oval the display. This problem is inherent in simple single-tuned channel filters and not at all bad as far as the terminal unit is concerned. To clean up the monitor scope traces W6FFC recommended a high-Q tuned circuit at the input to the scope; this is shown in fig. 1.

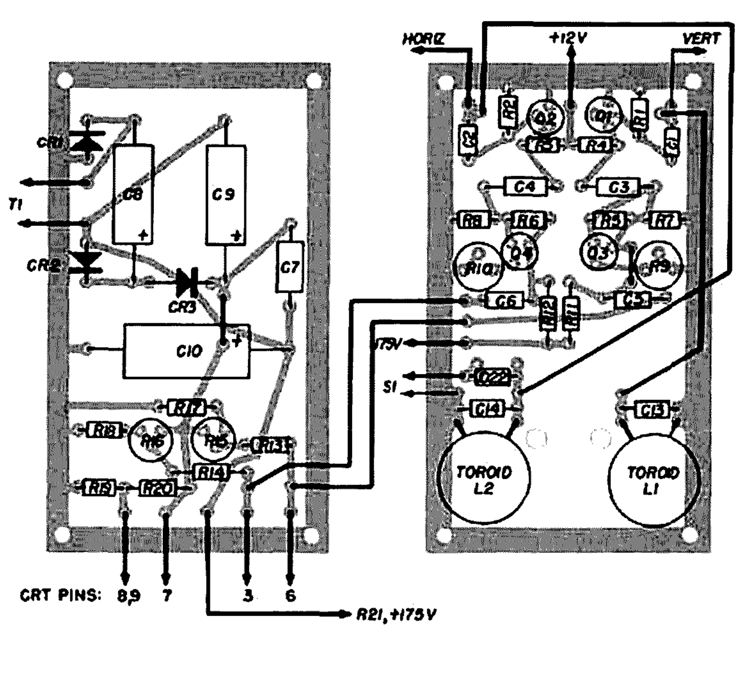

fig. 2. Printed-circuit board layout for the RTTY monitor scope.

Toroid L1 is tuned to 2125 Hz; L2 is tuned to 2975 Hz with S1 open. Close S1 and pad the 0.022-μF capacitor to tune L2 to 2295 Hz. These simple circuits are tuned so that maximum deflection of the scope trace coincides with maximum deflection of the ST-6 tuning meter.

construction

The RTTY monitor oscilloscope circuitry is built on two 3 x 6-inch printed-circuit boards. Circuit-board layout is shown in fig. 2. If you look at the boards carefully you will see that I have drilled two holes through the amplifier board so the centering controls (R14 and R15) on the power-supply board can be easily adjusted.

If you want, these controls could be located on the front panel, although they seldom require attention after the circuit has been initially set up. I put the focus control (R21) and intensity control (R22) on the front panel of my unit.

The power transformer I used for the 12-volt supply is the same as that used in the ST-5 and ST-6. However, use a separate transformer for the monitor scope; don't try to make the terminal-unit transformer do double duty.

For potentiometers R15 and R16 I used Ohmite type RV6NAV that I obtained surplus. You can use the Mallory MTC254L1 instead, but be careful because the whole potentiometer frame will be *hot* to the tune of 500 volts.

summary

This monitor scope is a nice companion to the Mainline ST-6 RTTY terminal unit. At my station I included it in the same cabinet as the ST-6. Performance is excellent, and the scope traces are clean and easy to read.

references

1. Irvin M. Hoff, "The Mainline ST-5 RTTY Demodulator," *ham radio,* September, 1970, page 14.

2. Irvin M. Hoff, "The Mainline ST-6 RTTY Demodulator," *ham radio,* January, 1971, page 6.

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resistor performance at high frequencies

A comparison of the high-frequency response of solid carbon-composition and metal-film resistors

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Most amateurs are aware of what happens to a capacitor as the operating frequency is increased – it becomes series resonant at the frequency where its internal inductance resonates with its capacitance. However, many amateurs may not realize that carbon resistors, as well as metal-film types, show resistive and reactive changes as the frequency is varied from dc to vhf.

Since resistors are the most common components used in electronic equipment it is helpful to know as much as possible about this *so-called* passive element. You may find that those resistors are not nearly as "passive" as you may think – under the right circumstances they may even exhibit a rather capricious nature.

In this article I will discuss the high-frequency characteristics of resistors. I will not cover wirewound types because of their high inherent inductance, even "non-inductive" types.

Before looking at high-frequency resistor performance we should briefly review the construction of both *composition* and *metal-film* types and the general characteristics of each. The *composition* resistor consists of a mixture of resistive material and a dielectric binder that is molded into a cylindrical shape. *Metal-film* resistors are composed of a resistive film deposited on, or inside, an insulating ceramic cylinder.

These two resistor types differ from each other in size, resistance range, cost, power dissipation and general characteristics. One type may be better than the other for particular purposes but neither type has all the best characteristics. Therefore, resistor choice depends upon the circuit requirements, the environment in which it must operate and many other factors.

composition resistors

Generally speaking, solid composition resistors have poorer stability since their resistance is a function of body temperature, circuit voltage, moisture content and previous history.

The high-frequency characteristics of composition resistors are good, but not quite as good as film types. Composition resistors generate a noise voltage when current flows through them; this is inherent in the construction of the resistor and is much greater than in any other type. Therefore, composition types are not recommended for circuits handling low-level signals.

The reliability of composition resistors is good since they seldom open up unless badly overloaded; however, they do change value by several percent with changing operating conditions. For example, they should not be used in voltage dividers where accuracy and voltage

stability is required. Also, they shouldn't be used where a long-term permanent resistance change of ± 10 percent or more cannot be tolerated.

In addition, composition resistors should be derated 50 percent* to increase life and stability in moderate temperatures, and even more than this for operation in high ambient temperatures.

film. However, exposure to moisture may seriously affect resistance if the element is not well protected by its casing.

The metal-film resistor should also be protected to reduce the possibility of physical damage. The temperature coefficient of resistance is extremely low (that is, the resistance change with change in operating temperature).

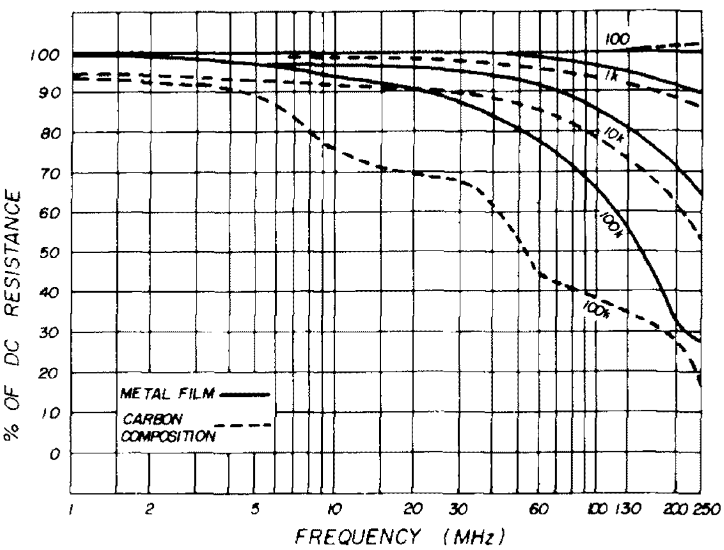

fig. 2. Resistance of both metal-film and carbon-composition resistors decreases with frequency. Larger values are affected more than smaller values.

metal-film resistors

Metal film type resistors generate considerably less noise voltage than composition types since the conducting paths are more homogeneous. Stability is high, environmental changes have little effect, and they have high reliability. Also, the combined effects of climate and operation will not change initial resistance by more than about 3 percent under normal operating extremes.

The maximum full-power operating temperature of metal-film resistors may be allowed to reach 70 to 100 degrees C. Skin effects are negligible since the entire resistance path is made up of a surface

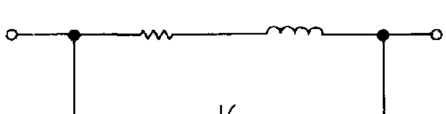

fig. 1. Simple equivalent circuit of a resistor includes series inductance and shunt capacitance.

high-frequency operation

A resistor that can be assumed to exhibit only resistance at low frequencies develops inductance and shunt capacitance as frequency is increased. See **fig. 1** for a simplified equivalent circuit of an ordinary resistor at high frequencies. In a well constructed composition or film resistor the inductance is due solely to the connecting leads.

Shunt capacitance is due to the capacitance between end caps (or the lead terminations) and the shunt capacitances formed by the conducting particles which are held in contact by the dielectric binder. The solid composition resistor naturally has a greater number of such contacts than the film type resistor.

resistance/reactance measurements

I have made a series of measurements on a set of good quality carbon composi-

*A ½-watt resistor should be used to dissipate ¼-watt maximum.

tion resistors and an equal number of quality metal-film types,* having values of 100, 1000, 10k and 100k ohms.

A Boonton RX Meter, model 250A, was used for the measurements. This is a wide frequency range impedance meter that is designed to give an accurate reading of the equivalent parallel resistance and parallel reactance of two-terminal networks or components. Two calibrated dials, labeled R and C, are used to balance a bridge.

The equivalent parallel resistance is then read from the R dial. The positive or negative resonating capacitance of the resistor (± pF) is indicated by the C dial.

high-frequency characteristics

All resistor leads were clipped to the same length (3/8 inch) before making any measurements. The first measurement was for dc resistance – all resistors were within the specified tolerances (±5% for the composition types and ±2% for the metal-film resistors).

Graphs of resistance and capacitance versus frequency for the four resistor values are shown in figs. 2 and 3. It can be seen that the resistance of the higher resistance units decreases at higher frequencies, with the metal-film types appearing to be slightly better performers.

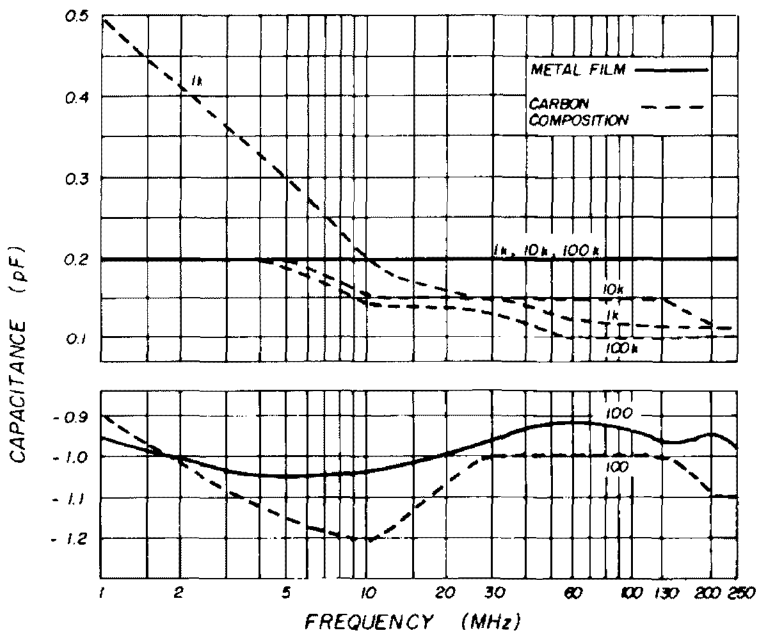

fig. 3. Shunt capacitance of carbon-composition resistors changes much more with increasing frequency than the shunt capacitance of metal-film types.

This drop-off in resistance with frequency was first discovered by Boella and is called the "Boella effect." Howe[1] suggested a transmission-line theory to describe this behavior, and this is generally considered correct.

At high frequencies the resistor looks like a network of resistances and capacitances according to Howe, so the resistance reduction with increasing frequency is due mainly to the shunting effect of distributed capacitance in the resistor (see fig. 4).

For the best high-frequency performance the controlling conditions are geometry (size and shape of the resistor) and minimum dielectric losses. It has

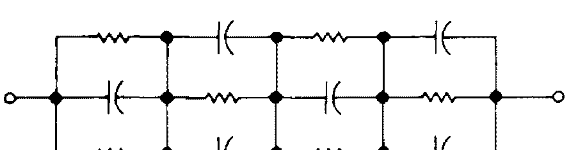

fig. 4. Realistic equivalent circuit of a resistor at high frequencies actually looks like an RC network.

*The composition resistors were Allen-Bradley type CB, ¼ watt, 5% tolerance. The metal-film resistors were Corning Glass type C4, ¼ watt, 2% tolerance.

been proven that the smaller the diameter of the resistor (minimum cross-sectional area), the better will be its high-frequency response, all other things being equal.

The small-diameter resistor will have fewer contacts to contribute capacitance than a larger unit. Generally the terminals and lead terminations will also be smaller and therefore contribute less capacitance. The dielectric losses are kept low by a good choice of base material. If binders are used their total mass should be kept to a minimum.

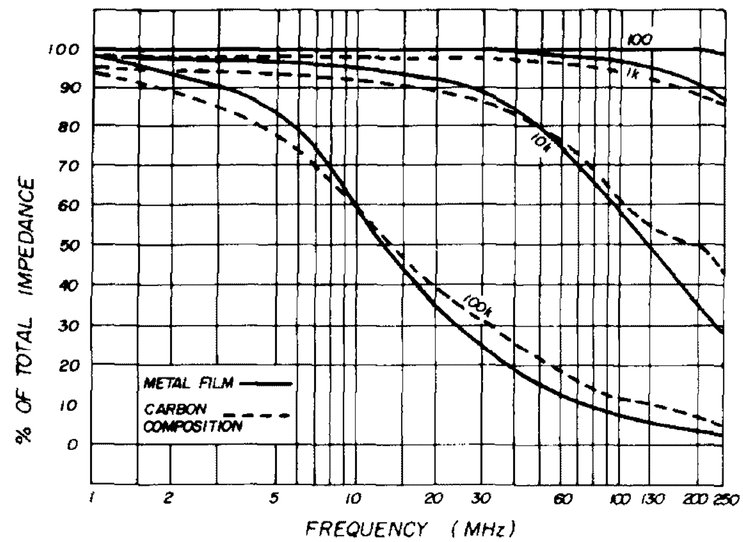

fig. 5. Total resistor impedance decreases with frequency. This is especially apparent with the larger values. Resistor impedance is a much more meaningful measurement than simple ac resistance.

total impedance

The total impedance of a resistor with shunt capacitance appears to be a more meaningful indicator of resistor performance at high frequencies than the basic ac resistance measurement.* A comparison of the curves in **fig. 5** indicates that very little difference exists between the two types of resistors at high frequencies if the basis of comparison is total impedance alone.

*The total impedance of the resistor is

$$Z = \frac{R_p X_p}{\sqrt{R_p^2 + X_p^2}}$$

where R_p is the equivalent parallel resistance and X_p is the equivalent parallel reactance (see **fig. 2**). X_p may be calculated for each test frequency using the standard reactance formula, $X = 1/2\pi fC$). The value of C is found in **fig. 3**.

conclusions

If you consider other criteria, such as stability, environment, inherent noise, high reliability, susceptibility to physical damage or cost, then one type or the other should be used for that reason and the high-frequency response should not generally be a major consideration.

If, however, high-frequency response is important, you should use the smallest physical size possible consistent with good design practice, e. g., if a 1/8-watt resistor will do the job safely, use it rather than a 1/4-watt or larger value.

The resistor leads and interconnecting wires should be as short as feasible and resistor placement should be chosen with care. Resistor capacitance to ground, for example, may greatly increase shunt capacitance and accentuate high-frequency roll-off.

references

1. G. W. O. Howe, *Wireless Engineer,* 12, 291 (1935); 12, 413 (1935); 17, 471 (1940).
2. Charles L. Wellard, "Resistance and Resistors," McGraw-Hill, New York, 1960.
3. Keith Henney, Craig Walsh, "Electronic Components Handbook," Chapter III, Technical Writing Service; McGraw-Hill, New York, 1957.
4. John F. Blackburn, "Components Handbook," Volume 17, MIT Radiation Laboratory Series, McGraw-Hill, New York, 1949.

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modified inverted-V antenna

This modified inverted-V antenna provides complete multiband operation from 40 through 10

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Many amateurs I have worked have expressed interest in my rather unconventional multiband antenna. I call it a modified inverted-V because it began as an inverted-V for 40 meters. However, it has been modified considerably and now provides good performance on all bands from 40 through 10 meters.

For those amateurs who are already using an inverted-V antenna, this design may give them some ideas for covering other bands. For the amateur with limited space this antenna offers good efficiency with minimum size.

The multiband inverted-V antenna shown in **fig. 1** offers many advantages, including small physical size, relatively low height, broadband response with low swr and requires no traps or tuning devices. In addition, it appears to be nearly omnidirectional.

The vswr curves plotted in **fig. 2** were measured with a Knight P2 swr bridge. As you can see, the antenna is cut for phone

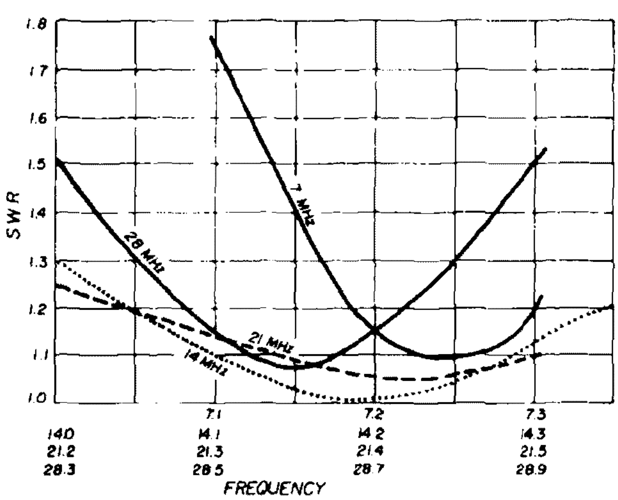
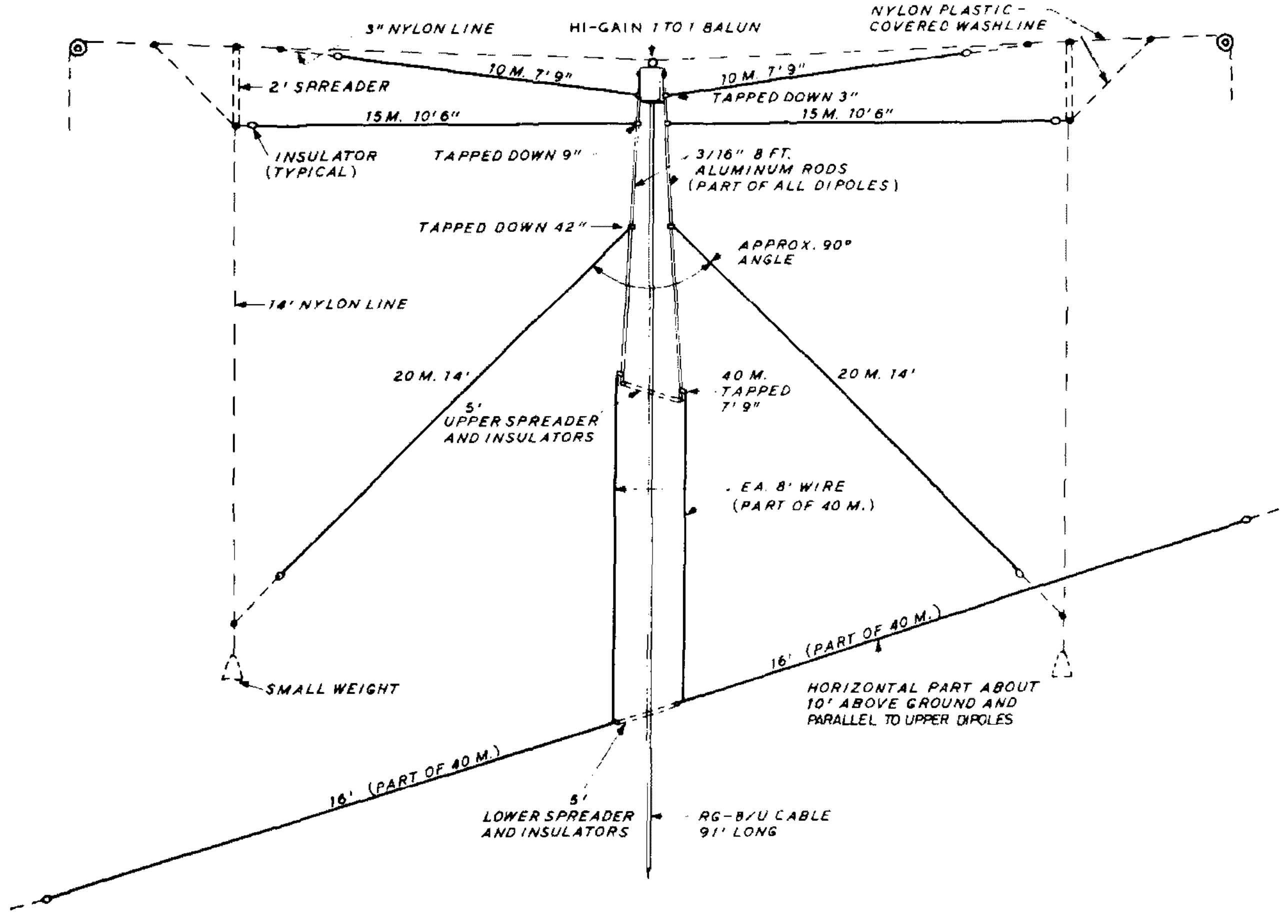
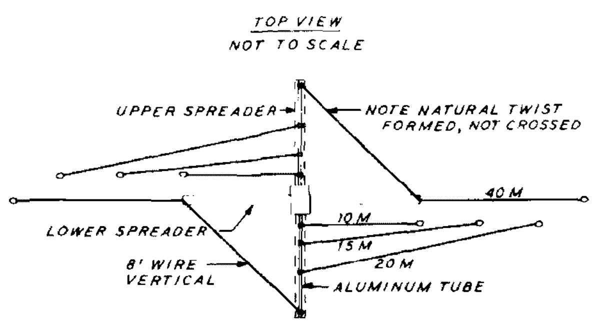


fig. 2. Swr performance of the W2KTW modified inverted-V.





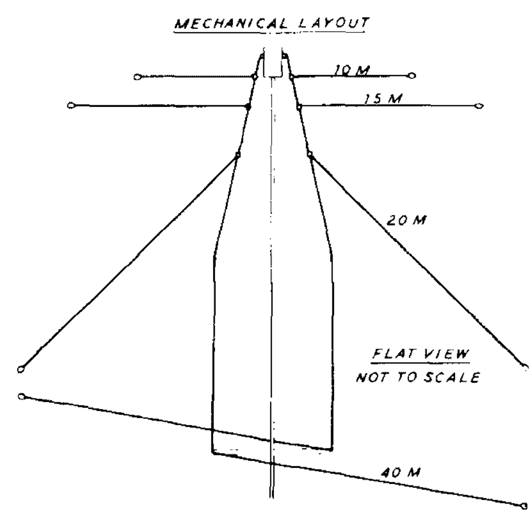


fig. 1. Modified inverted-V antenna provides good operation on 40, 20, 15 and 10 meters.

portions of the amateur bands. Although my modified inverted-V has the dimensions given in **fig. 1**, at other locations it will probably be necessary to trim each of the sections to resonate at the desired frequency.

On 20 meters, where the antenna appears as a more conventional inverted-V it was noted that decreasing the angle between the elements raised the resonant frequency.

On 40 meters it was noted that the horizontal part of the elements must be 180° away from the higher-frequency elements. Also of importance is the fact that the vertical support provides more of a twist than a transposition.

With this antenna I have obtained optimum loading on all bands with my TR-4 transceiver. In the future I hope to devise a way of including a 75-meter antenna within the limited space I have available.

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series-tuned pi networks

An output tuning network that increases the upper frequency limit of many power-amplifier circuits

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One of the problems confronting the rf power amplifier designer in the upper hf and the lower vhf range from 20 to 150 MHz has been the achievement of tank-circuit constants that provide normal operating Q and decent power transfer. For years the bandswitched pi-network has been the accepted tank circuit for most amateur multiband rf power amplifiers. Unfortunately, this has precluded some otherwise excellent inexpensive rf power amplifier tubes from use above 15 meters because the tube plate output capacitance, added to the minimum capacitance of the input tuning capacitor and other stray circuit capacitance, has made it impossible to achieve a reasonable operating tank circuit Q.

When high-C, low-L tanks are tried with these tubes the results are generally unsatisfactory due to high circulating currents in the tank inductor which result in heating and loss of efficiency.

Attempts have been made to overcome this deficiency by eliminating the pi-network input tuning capacitor; to tune the tank the inductance is adjusted by a moveable slug or shorted turn. Usually this results in complicated or unwieldy mechanical assemblies that are limited to single-band amplifiers, and power losses are inherent in the slug or short-circuit turn.[1,2] For example, one of the problems that detracted from the use of parallel 813s in bandswitched pi-network amplifiers was that of obtaining reasonable operating Q on 10, 15 and occasionally, 20 meters. Reasonable Q was particularly unattainable on 10 meters due to the high plate output capacitance of two 813s in parallel (28 pF). Therefore, use of two 813s on ten meters was virtually restricted to push-pull, plug-in coil, link-coupled amplifiers or to single-band configurations. The single 2E26, or two 6146s, on two meters is normally restricted to compromise LC tank circuits which use link output coupling. Since the

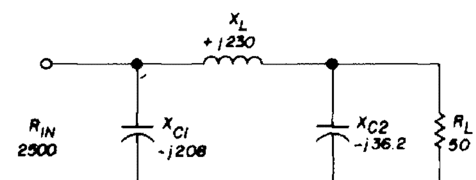


fig. 1. Basic pi network. Component values are discussed in text.

link has to be series tuned the transmitter ends up with as many tuning controls as a pi-network, but without its advantages.

tank circuit design

The operating Q of most plate tank circuits is chosen in the range of 10 to 20. In the pi-network the operating Q is a function of tube plate resistance and capacitance at the input to the network. Varying either of these parameters affects the component values of the pi-network.

A careful review of the reference material will reveal that to maintain a fixed value of operating Q as plate voltage and current (and therefore plate resistance) are changed, the values of C and L must be changed. For a given plate voltage/current ratio the Q will vary directly as the tank capacitance; ie, doubling tank capacitance doubles the Q.

For a given value of Q the input tuning capacitor in a pi-network can be larger with a low plate voltage/current ratio than with a high plate voltage/current ratio.

To achieve an operating Q of 12 in a pi-network with a plate voltage/current ratio of 5 (1500 volts at 300 mA or 2000 volts at 400 mA), the reactance of the input tuning capacitor is expressed by

$$X_C = \frac{R_p}{Q} \tag{1}$$

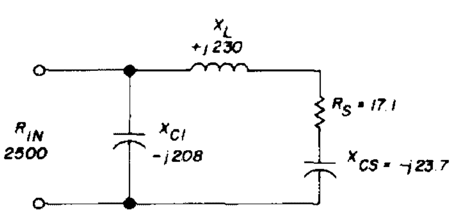
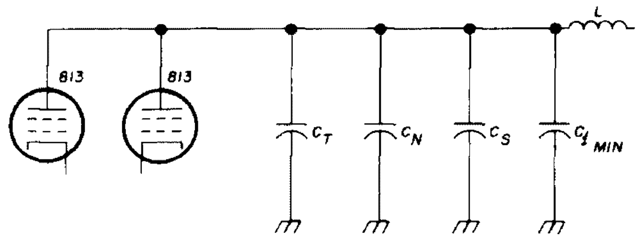


fig. 2. Series-tuned pi network that is electrically equivalent to the circuit in fig. 1.

where R_p is the tube plate resistance and is equal approximately to $E_p/2I_p$. For the case of the amplifier operating at 1500 volts and 300 mA with a Q of 12 the input tuning capacitor should exhibit 208 ohms capacitive reactance. This represents a 27.3-pF capacitor at 28 MHz. The other pi-network values may be determined from the formulas given in the ARRL Handbook.[3] For this case the reactance of the output capacitor is 36.2 ohms (157 pF at 28 MHz). The required inductance is 1.31 μH at 28 MHz (230 ohms inductive reactance).



type of capacitance	bandswitched amplifier	single-band amplifier
C_{tube}	28 pF	28 pF
$C_{neutralizing}$	6 pF	6 pF
C_{stray}	20 pF	6 pF
C_1 minimum	25 pF	15 pF
Total C	79 pF	55 pF
X_C	72 ohms	103 ohms
Minimum realizable Q for 2500-ohm plate load	34.7	19.25

fig. 3. Estimates of input tuning capacitance for bandswitched and single-band rf power amplifiers.

The pi-network has the basic form shown in fig. 1 with the output load R_L in parallel with the loading capacitor X_{C2}. The reactance values are those given above. This basic circuit arrangement can be converted into the series-tuned configuration in fig. 2 by finding the equivalent series impedance of the two output components:

$$Z_S = \frac{R_L X_{C2}}{R_L + X_{C2}}$$

For the circuit constants in fig. 1, the equivalent series impedance is 17.1 - j23.7 ohms.

At the resonant frequency the network looks like a pure resistance at the

input terminals R_{in}. Although this matches the plate resistance of a typical pair of 813s it is impossible to use this circuit on 10 meters because the tube output capacitance and stray circuit capacitance are greater than the minimum value of C1. This is the limiting factor in tank circuit Q at the highest frequency of operation.

It becomes extremely important to arrive at a circuit configuration and operating voltage/current ratio that will allow acceptable Q on ten meters. Fig. 3 presents an estimate of total pi-network input circuit capacitance in two actual parallel 813 amplifiers. As can be deduced from the data, on 10 meters the tank circuit of the bandswitched amplifier will probably get hot enough to boil water.

series pi-network

Since capacitive reactance cancels a like amount of inductive reactance in a series combination of C and L, a method of varying L in a circuit is suggested. If the inductive reactance of L has a certain value, and a capacitive reactance of some smaller value is placed in series with it, the inductance of L will be effectively reduced. If the capacitive reactance is variable, then L can be varied just as effectively as if it were a roller inductor or tapped coil.

If the input tuning capacitor of the pi-network (C1) is insulated from ground and placed in series with the inductor the minimum capacitance limitation is virtually eliminated, reducing the total fixed capacitance to no more than 40 pF (see fig. 4).

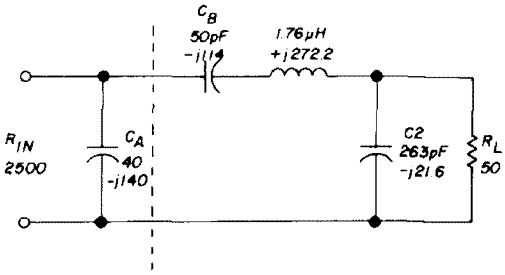


fig. 6. Input impedance of circuit in fig. 5 may be increased to 2500 ohms by changing component values as shown here.

If the reactances to the right of the dotted line in fig. 4B are added together (since they are in series) the total series impedance is 17.1 + j140. Thus the circuit of fig. 4 may be simplified to the form shown in fig 5. This parallel resonant circuit is exactly the same result you would obtain if you reduced the standard pi-network to its most simple equivalent form.

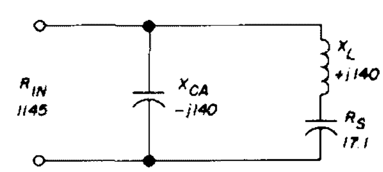


fig. 5. Electrical equivalent of the series-tuned circuit in fig. 5.

At resonance the input resistance to the parallel tuned circuit is

$$R_{in} = \frac{X^2}{R_S} \quad \frac{(140)^2}{17.1} = 1145 \text{ ohms}$$

Tank circuit Q, from *eq.* 1, is 17.85. An impedance match can be obtained by increasing the value of the tank coil fig. 4A) to 1.76 μH and decreasing the series tuning capacitor to 50 pF. The network is matched to the load by increasing C2 to

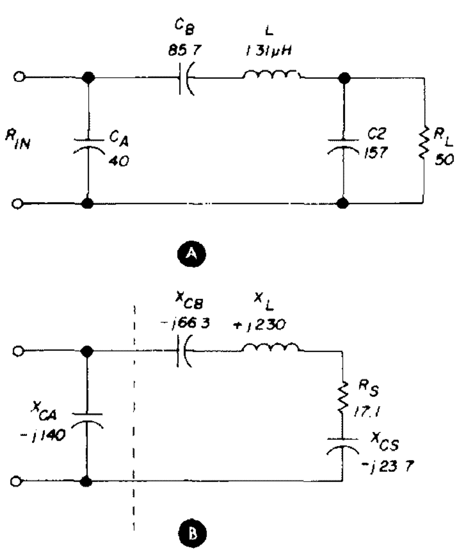
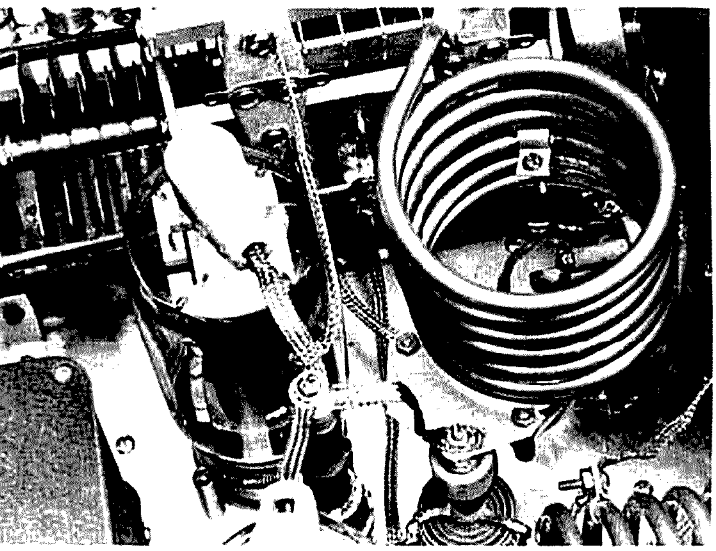


fig. 4. Tuning capacitor may be placed in series with the inductor (A). Equivalent reactance values are shown in (B). This circuit is electrically equivalent to circuit in fig. 5.

263 pF. The new component values and their reactances are shown in fig. 6.

The equivalent series impedance of R_L and C2 is 7.85-j18.2 ohms. Total reactance to the right of the dotted line is still +j140. However, R_S is now 7.85 ohms, so the input resistance, from *eq. 2* is

$$R_{in} = \frac{(140)^2}{7.85} = 2500 \text{ ohms}$$



Bandswitched parallel 813 rf power amplifier using the series-tuned pi network.

The operating Q is 17.85; this value is perfectly acceptable.

In cases where there is enough tube output capacitance plus nominal stray capacitance to equate to the normal input tuning capacitor in a regular pi network the series-tuned configuration warrants prime consideration as an efficient tuning scheme. The series-tuned pi-network has the same number of components as a regular pi-network, and they all have reasonable, attainable values.

There is actually better harmonic attenuation with the series-pi than with the regular pi for a given situation due to the lower shunt reactance to ground on the output side of the network. The network will match a broad range of load resistances for a given Q and plate resistance.

If a Q of 20 is considered the top acceptable limit, the maximum fixed tube output plus nominal stray capacitance for a given plate resistance that can be accommodated with the series-pi network is given in table 1.

table 1. Maximum tube output capacitance and stray capacitance (Ca) that can be accommodated by the series-pi-network for a Q of 20.

Rp (ohms)	Xca (ohms)	Ca (pF) 28 MHz	50 MHz	144 MHz
1000	50	114	64	22.1
2000	100	57	32	11.1
2500	125	45.5	25.5	8.85
3000	150	38	21.3	7.4
3500	175	29.2	18.4	6.3

bandswitching

Most of the input capacitance in a normal tank circuit, other than tube output capacitance, is a result of long leads associated with bandswitch circuitry. The preceding network analysis was based on the absence of bandswitch circuitry at the input to the network thereby keeping stray capacitance to an absolute minimum.

You can build a bandswitching amplifier with the series-tuned pi-network if you place the switch on the output side of the network as shown in fig. 7. Since there are no particular restrictions on the input tuning capacitance on 40 and 80 meters, a standard pi-network is used on those bands. The lead from the tube plates through the coupling capacitor to the switch adds only a few pF to the stray capacitance and can be virtually ignored.

Additional inductors are switched into the circuit for 15 and 20 meters. The photograph shows the installation of the 10/15-meter inductor. A section of the 20-meter coil can be seen in the lower right-hand corner of the photo.

alternate series-pi

For even greater reductions in stray capacitance the series tuning capacitor is placed between the inductor and the load as shown in fig. 8. With this arrangement the stray capacitance is on the output side of the network where it becomes a

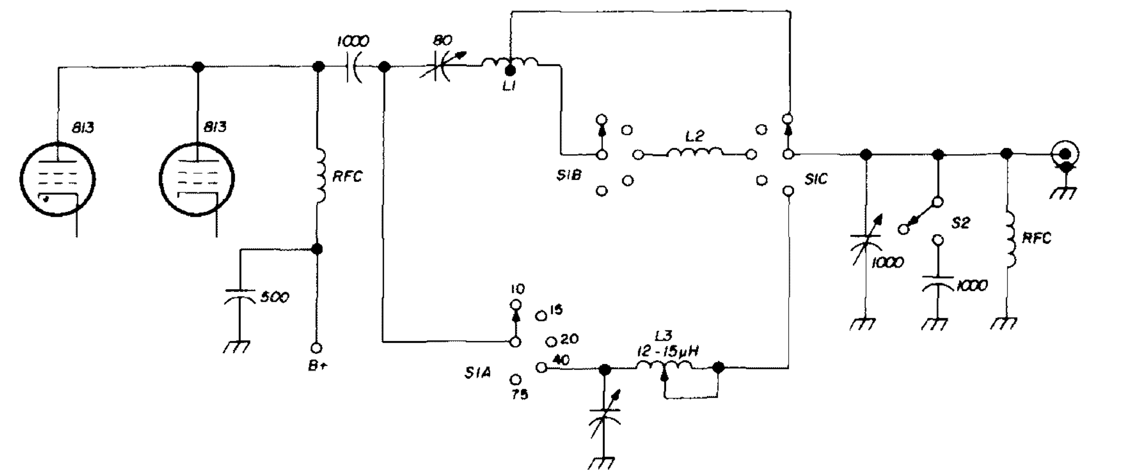
very small fraction of the total shunt capacitance to ground.

To verify the alternate approach, a two-meter 2E26 transmitter was altered to the series-tuned pi-network configuration in **fig. 8.** The plate output capacitance of a single 2E26 is 7 pF, exactly the magnitude of tank capacitance required for a Q of 12 with a plate voltage of 300 and plate current of 75 mA.

The original two-meter tank circuit was a push-pull arrangement with a series-tuned link-coupled output. **Fig. 9A** shows the original circuit; the photograph shows the original circuit in operation with a vtvm indicating *relative* rf voltage across a 50-ohm dummy load.

The same circuit components were reconnected into the arrangement of **fig. 8.** Fig. **9B** illustrates the improvement in efficiency as demonstrated by the relative output rf voltage reading. Every attempt was made to keep all test conditions with the exception of the tank circuit the same for both measurements.

In addition to the improved efficiency of the series-tuned tank circuit there were other advantages. Tuning was much smoother than the original, and there were no rf feedback problems when the tank circuit was slightly off resonance. The final results with the series-tuned tank were well worth the 30 minutes it took to move the coil and re-solder a few connections.



L1 6 turns ¼" copper tubing, 2 turns per inch, 3" OD; 10-meter tap 2 turns from output end

L2 6 turns ¼" copper tubing, 2 turns per inch, 3" O. D.

L3 12 — 15 μH roller inductor, 500- to 1000-watt rating

fig. 7. Using the series-tuned pi network in a multiband band-switched power amplifier.

design procedure

The following step-by-step procedure will help you design a series-tuned pi network for your own particular requirements. **Steps 1** through **4** will indicate whether the series pi is a practical solution to your design problem.

1. Determine tube operating conditions to achieve the lowest possible tube plate resistance consistent with power input requirements

$$R_p = \frac{E_p}{2\ I_p}$$

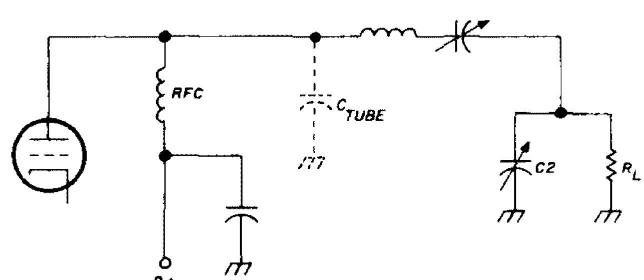


fig. 8. Effects of stray capacitance may be further reduced by placing the series tuning capacitor between the inductor and load.

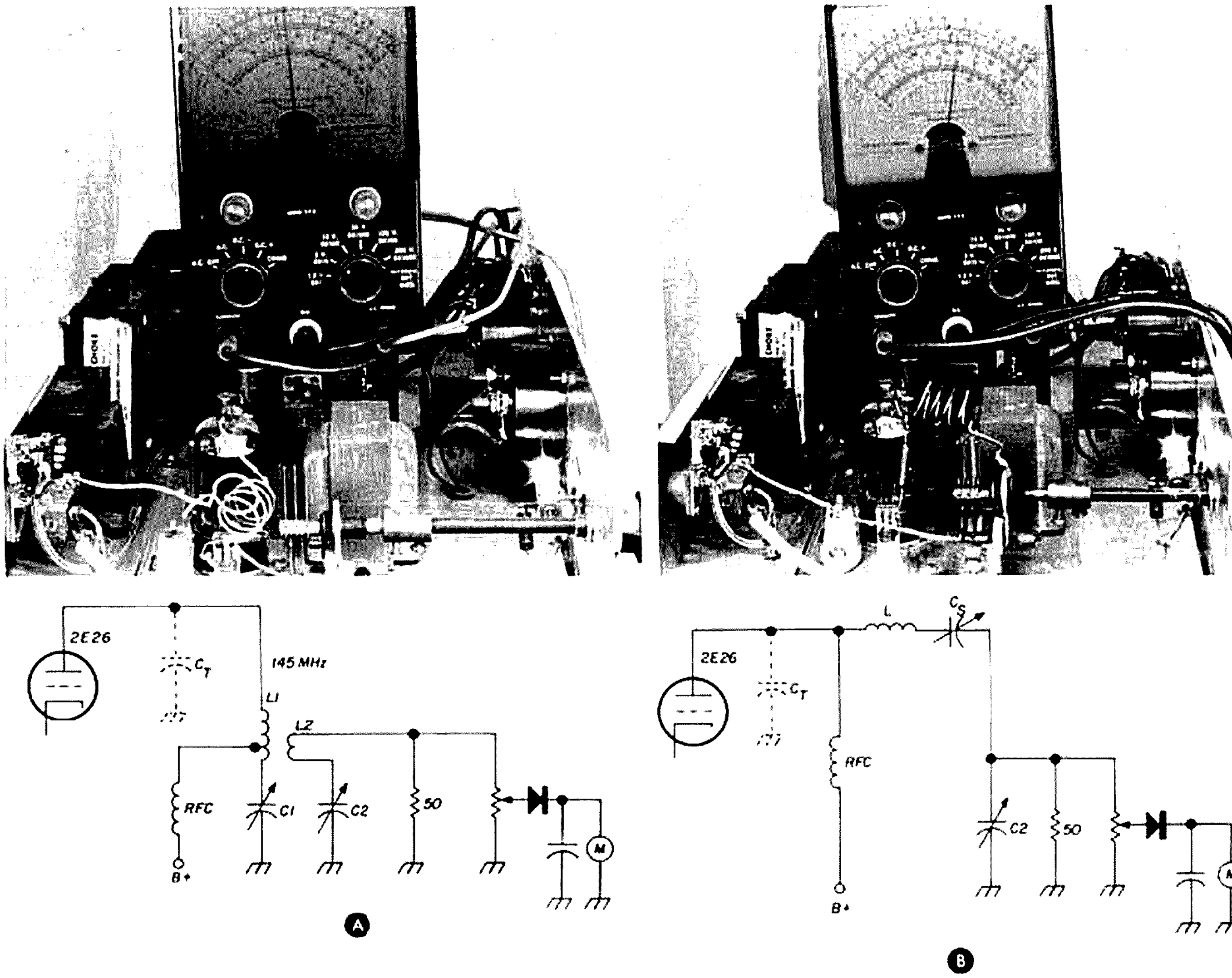


fig. 9. Original two-meter power-amplifier circuit is shown in (A). Series-tuned pi network in (B) provides increased rf output.

2. Determine total fixed input capacitance to the network by adding tube output capacitance to approximately 15 pF stray capacitance

$$C_{in} = C_{tube} + C_{stray}$$

3. Convert C_{in} to its equivalent reactance

$$X_c = \frac{1}{2\pi f C_{in}}$$

4. Compute Q

$$Q = \frac{R_p}{X_{c\ in}}$$

If Q is not less than 20, re-examine plate resistance R_p and attempt to find a new set of operating conditions to arrive at a lower value.

5. X_c is equal to X_L', that part of X_L remaining after the reactance of both the series tuning capacitor (X_{cb}) and equivalent series loading capacitor (X_{cs}) are subtracted from the total inductive reactance

$$X_c = X'_L$$

6. Compute the required series output load resistance R_s

$$R_s = \frac{(X'_L)^2}{R_p}$$

7. Knowing R_s, the corresponding X_{cs} portion of the series equivalent load impedance can be selected from fig. 10. This graph is calculated for *only* a 50-ohm antenna at 28 MHz.

8. Select a reasonable value for the series tuning capacitor C_B and calcu-

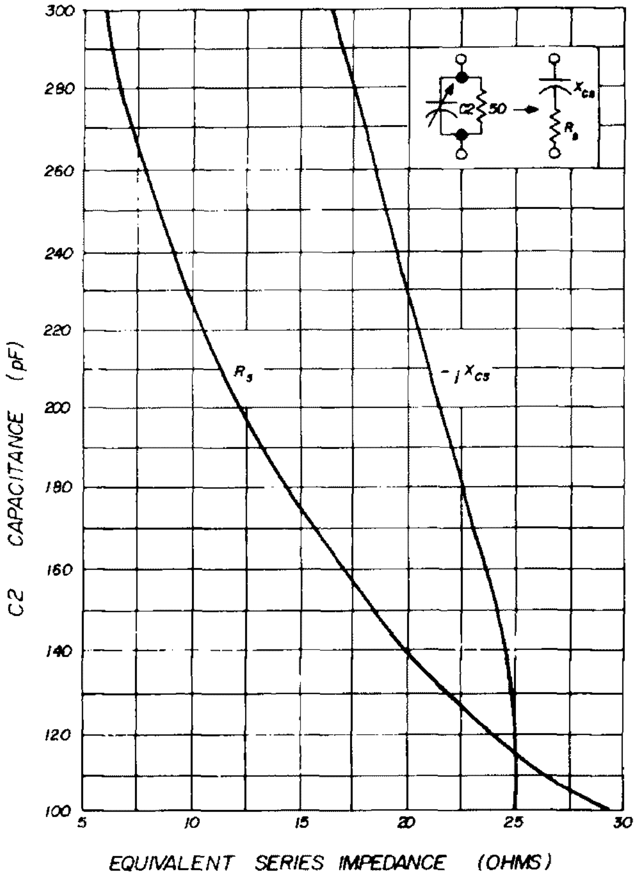


fig. 10. Graph for converting pi network parallel impedance to equivalent series impedance. (Capacitor C2 in parallel with 50-ohm load at 28 MHz.)

late its reactance using the formula in **step 3.**

A value of around 100 ohms for X_{CB} is suggested since this results in about 57 pF at 28 MHz for C_B. This is a reasonable midrange value for a tuning capacitor that will cover 10 through 20 meters.

9. Determine X_L by adding the absolute values of X_L', X_{CB} and X_{CS} (ignore the plus/minus j factors).

$$X_L = X_L' + X_{CB} + X_{CS}$$

10. Calculate the value of the tank inductor from the reactance value determined in **step 9**

$$L = \frac{X_L}{2\pi f}$$

11. Determine the value of the output loading capacitor which results in the equivalent series impedance found in **step 7.**

$$X_{C2} = \frac{R_S^2 + X_s^2}{X_S}$$

The capacitance value of C2 is calculated with the reactance formula given in **step 3.**

This completes the design of the series-tuned pi-network. Double check all calculations.

summary

The results of using series-tuned pi-networks has been very gratifying. Tuning and loading with the series circuit is no different than with a standard pi-network. Dial settings for tuning capacitance, inductance and loading capacitance for a resistive 50-ohm load are right on the calculated optimum values.

The series-tuned pi-network should permit the use of inexpensive power tubes in bandswitched linear cathode-driven grounded-grid amplifiers. The problem of plotting a set of design curves for the series tuned pi-network (similar to the pi-network charts in the ARRL Amateur's Handbook) will be left to an industrious engineering student with access to a digital computer.

references

1. William Orr, W6SAI, "Inductively-Tuned High-Frequency Tank Circuits," *ham radio,* July, 1970, page 6.

2. Robert Sutherland, W6UOV, "Two-Kilowatt Six-Meter Linear," *ham radio,* February, 1971, page 16.

3. "The Radio Amateur's Handbook," A. R. R. L., 46th edition, 1969, "R. F. Circuits," pages 41-49.

4. Truman S. Gray, "Applied Electronics," John Wiley & Sons, 5th edition, 1957, Chapter IX, "Tuned Circuits in Amplifiers," page 547.

5. Frank Jones, W6AJF, "High-Frequency Operation of Parallel Tubes," *CQ,* September, 1946, page 25.

6. Douglas Blakeslee, W1KLK, Carl Smith, W1ETU, "Some Notes on the Design and Construction of Grounded-Grid Linear Amplifiers," *QST,* December, 1970, page 22.

ham radio

power in reflected waves

A discussion of power in reflected waves along a transmission line, and its effect upon equipment operation

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Discussions of reflected waves on radio-frequency transmission lines sometimes present the concept that energy in these waves is absorbed by the transmitter; this would adversely affect the operation of the output tuned circuit, or the output tube, or both, by generating heat therein. Of course, if this concept were correct, the result would be the transfer of less rf energy from the transmitter to the load at the far end of the line, usually an antenna. The loss of energy in such case would depend on the magnitude of the reflected wave in relation to that of the forward or incident wave (the swr) and on the proportion of energy in the reflected wave that is absorbed by the transmitter.

As to the proportion of the energy in the reflected wave that is absorbed by the transmitter, opinion varies. Some say that this proportion may range as widely as that of the forward wave at the load end, all the way from zero to 100%, depending on the degree of match between the impedance of the output tuned circuit and the input inpedance of the line.

If this match is good, the absorption would be large according to this concept, and nearly all of the energy of the reflected wave would be absorbed. This seems to be a strange situation because the adjustment of the transmitter output circuit is to make the match to the line as good as possible for best forward transfer, whereby most of the energy in the reflected wave would then be absorbed and wasted! (It should be remembered of course that adjustments at the transmitter have no effect on the swr near the load; this is determined solely by the degree of match between the line and the load.)

absorption concept

It may be useful to examine this concept of energy absorption at the transmitter to see whether it stands up under such examination, and to see if there is an acceptable alternative to it. (Hereafter I will refer to this as the *absorption concept.)* To do this, it will be necessary to consider specific transmission lines and impedance-matching devices such as tuned circuits or their equivalents.

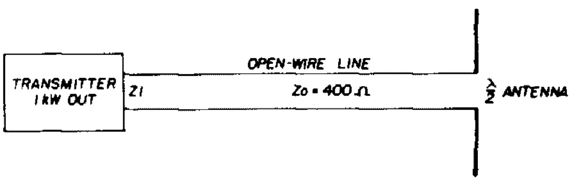

fig. 1. Feeding a half-wave antenna with open-wire transmission line results in a high standing-wave ratio. However, efficiency of this system is very good.

At first we will consider these as idealized circuit elements with no inherent losses, and later consider actual circuit elements with their small but unavoidable losses, to see what effect if any these losses may have on our conclusions. This will make it much easier to discuss the main issue: The possibly much larger losses that may result from high swr if the absorption concept is correct.

First, let us see whether the absorption concept leads to results that are found in practice. Consider **fig. 1**. There we have a transmitter with a power output of 1 kW with a tuned output circuit capable of being tuned to resonance and of providing a good impedance match between the plate resistance of the tube and the input impedance Z_i of the parallel-conductor open transmission line. This line, which is specified to have a characteristic impedance Z_o of 400 ohms, feeds the end of a half-wave antenna so there are standing waves on the line because of a large mis-match between the line and the antenna feed point.

The swr could be very high, well above 10, but we will assume that it has the reasonable value of 9. Because of these standing waves the line input impedance Z_i, as seen by the transmitter, may depart considerably from the nominal value of 400 ohms, depending on the electrical length of the line; it may be larger or smaller and it may exhibit reactance as well as resistance. Whatever this line input impedance may be we will specify quite reasonably that the transmitter output circuit provides a good match for it. Thus, according to the absorption concept the transmitter absorbs nearly all the energy in the reflected wave. This, calculated as usual, turns out to be represented by a power of 640 watts!

Many amateurs, including myself, have successfully operated amateur transmitters using similar feed lines which were, in all essential respects, equivalent to the arrangement shown in **fig. 1** without the slightest evidence that anything approaching 64% of the energy going into the transmission line was being returned to the transmitter, appearing as heat in the tuned circuit or tube or both. If even a fraction of that energy was coming back into the transmitter it would have been immediately evident to a knowledgeable amateur that something was very wrong.

Some amateurs have experienced overheating in amplifier tuned circuits or tubes; this can easily happen if the tuned circuits are not properly designed or built, and particularly if the coupling between the tuned circuit and the line is not sufficient to bring the loaded Q of the circuit down to the proper value, usually about 12. With insufficient coupling the circulating current in the tank circuit can rise to a high value and may damage the inductor or other parts. However, as such, this is not the fault of the high swr.

If the absorption concept were correct these parallel-conductor open lines with their high swr would give incessant trouble. In fact, these same open transmission lines are very efficient in spite of their high swr. Thus, the absorption concept is not confirmed by qualified experience.

power factor

Now consider fig. 2. This arrangement uses the same transmitter on the same frequency but with output circuits adjusted or modified to match a short 50-ohm coaxial cable which is connected to the input of an impedance-matching device. This device accurately matches the 50-ohm cable to whatever may be the input impedance Z_2 of the same open-wire transmission line as used before with the same swr of 9. (Z_2 may not equal Z_1 because the transmission line is not necessarily the same electrical length.)

The transmitter cannot now absorb any energy that may exist in the reflected wave in the open transmission line for there is no reflected wave in the 50-ohm cable connected to the transmitter since the 50-ohm line is perfectly matched at both ends.

Where does any such energy, if it exists, appear? It cannot disappear. Are we to believe that it must appear as heat in the impedance-matching device? That would be doing the designers and constructors of such devices a grave injustice because it is readily feasible to design and build such devices having a *very* much smaller loss than seems to be implied in our example.

The small residual loss in matching between lines can be reduced nearly to zero by using a suitable tuning stub or in some cases a quarter-wave linear transformer, in place of a network with lumped constants. We cannot reasonably believe that any considerable proportion of the calculated power in the reflected wave on the open-wire line is resulting in the dissipation of heat in our matching device. Then where is it?

The simple answer is that it never existed, and therefore, there is no problem of power loss or dissipation of heat to be solved. The only real power in the transmission line (barring the previously mentioned small inherent losses) is that which results in the dissipation of energy in the load. That is equal to the difference between the *indicated* values for forward and reflected power.

It is usually understood that the forward wave and the corresponding reflected wave are each made up of two associated waves, one of current and the other of voltage. The usual discussion proceeds to show how the forward voltage wave adds to and subtracts from the reflected voltage wave, taking phase as well as magnitude into account, to form a voltage standing wave.

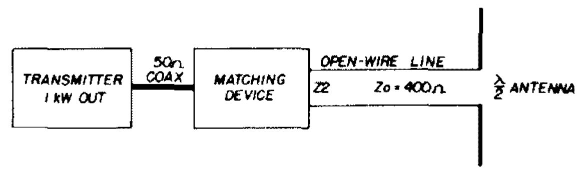

fig. 2 Antenna matching device provides good match to transmitter although there is still a high swr on the line to the antenna.

Similarly, it is shown how the forward and reflected current wave interact in the same fashion to form a current standing wave. What is not always made clear is that the two resultant standing waves (one of voltage and the other of current) have a phase difference of 90°. The importance of this fact becomes clear when it is recalled that in an ac circuit power is equal to voltage times current times the power factor:

$$P = EI \times \text{power factor}$$

Power factor is the cosine of the phase angle and the cosine of 90° is zero. Therefore, the power in standing waves along a transmission line is zero.

In the case of a lossless line (either shorted or open at the load end or ending in a pure reactance) reflection is complete. The two standing waves constitute the entirety of electric waves on the line. That is, the only current in the line is that in the standing wave of current and the only voltage on the line is that of the standing wave of voltage. As a result of the phase angle of 90° between the two powers in the line of zero.

The usual directional rf wattmeters do not take power factor into account; in the case of our lossless line with complete

reflection, the rf wattmeter will indicate equal forward and reflected "power," their difference being zero. An analogous situation exists if we connect an ideal (lossless) capacitor to an ac source. An ac voltmeter across the capacitor will indicate volts and an ac ammeter in series will indicate amperes, but there is no energy dissipation in the capacitor, and therefore, no power in the connections to it. The phase angle between the voltage and current waves in a capacitor is 90° so the power factor is zero.

Next consider a lossless transmission line terminated in a load that is purely resistive and equal to the characteristic impedance of the line. Under these conditions there is no reflection, no standing waves, and all the energy into the line appears as energy into the load. The power factor for the line in this case is 1.0. That is, the current and voltage waves coming from the transmitter to the load are in phase.

Another termination for our lossless line is one having a resistance differing from the characteristic impedance of the line or having some resistance and some reactance. such cases there will be some reflection but some energy will be dissipated in the load. The analysis of this situation is somewhat more complex but will be facilitated by considering that a *part* of the total line current is assignable to the standing wave of current and that a *part* of the voltage on the line is likewise assignable to the standing wave of voltage. These portions of the total line current and voltage are 90° out of phase as in the previously described cases and therefore do not represent power. The remaining current and voltage are in phase and represent real power which results in dissipation of energy in the load. Thus, the *only* power in the line is that which flows from the tranmitter toward the load.

The line as a whole will have a power factor at the input end of just the value needed to satisfy the relation

$$\text{power factor} = \cos\theta$$

where the angle O is the phase angle between the total line current and total line voltage measured at that point. This same statement holds even if the line has some losses. The only power in the transmission line at its input end is that power which accounts for line losses as well as for energy dissipated in the load.

conclusion

The concept of reflected "power" is a useful fiction to help us visualize the formation and nature of standing waves, but it can get us into trouble if taken too literally. If you insist on considering the reflected wave as real power then you must adopt another fiction, namely that when it gets back to the transmitter it is *completely* re-reflected toward the load. (The alternative is the absorption concept which I hope by now has been given a decent burial.) Complete re-reflection of power at the input end is impossible to accept since the necessary conditions of impedance mismatch are not present. This is perhaps an example of one untruth, or even two, never being enough!

Although open transmission lines were used in the previous discussion, exactly the same arguments apply to systems using coaxial lines and result in exactly the same conclusion: Power in a transmission line flows in one direction, from transmitter toward the load.

references

1. F. E. Terman, "Radio Engineering," McGraw-Hill, New York, 1947, chapter 4.
2. Woodrow Smith, "Antenna Manual," Editors and Engineers, Santa Barbara, 1948, chapter 2.
3. Byron Goodman, W1DX, "Losses in Feed Lines," *QST,* December, 1965.
4. R. C. Hills, "Some Reflections on Standing Waves," *RSGB Bulletin,* January, 1964.
5. P. M. Chamberlain, "Swr and Tank Coil Heating," *CQ,* August, 1966.
6. Carl Drumeller, W5JJ, "Effect of Mismatched Transmitter Loads," *ham radio,* September, 1969.
7. "Antenna Handbook," A. R. R. L., Newington, Connecticut, 1970, chapter 3.
8. "Antennas and Radio Propagation," Army Training Manual TM 11-666, 1953.
9. "Fundamentals of Electronics," NAVPERS 93400A-3, 1965, volume 3.

ham radio

swr bridge

An in-line directional bridge for monitoring standing wave ratio

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Although there are many swr indicator circuits and construction techniques, most are unnecessarily complicated. The one described here is simple and easy to make, requiring no special hardware. In addition it provides, on one meter, simultaneous indications of both forward and reflected voltage, leaving both hands free to adjust transmitter controls.

operation

The indicator **(fig. 1)** is somewhat more intricate than most, but its advantages outweigh its complications.* Its operation is based on the use of a zerocenter dc microammeter. The diodes are connected to give opposing polarities, unlike most circuits, so that FWD gives an indication to one side of the meter; REF to the other. When the switch is in FWD or REF position, operation is the same as with other bridges. In the center position, the voltages are combined in a resistive divider, and the meter gives an indication of the relative predominance of either forward or reflected voltage. Be sure to connect the switch so that forward voltage will be read in the same direction as the switch when it is turned to FWD.

construction

The pickup line is the result of a search for an easy-to-make, effective, and small unit. It has been described before.[2] It uses the inner conductor of a piece of RG 58/U or RG 11/U.

The pickup line is constructed as shown in **fig. 2.** Tin the edges of the hole in the braid to prevent shorting to resistor R_T, then slip the inner part of the coax inside the braid. Solder the ends of the

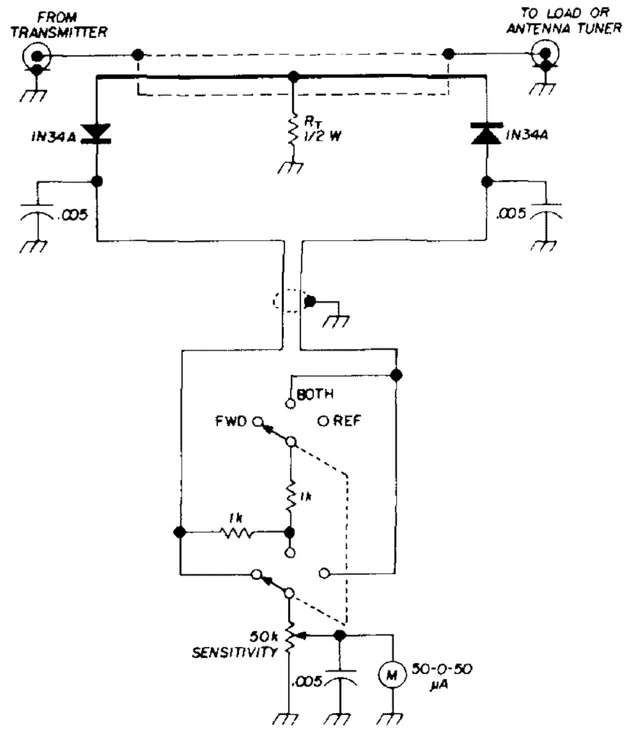


fig. 1. Schematic of bridge and indicator. Resistor R_T is equal to the value in ohms of the load impedance (see text).

*This design is similar to the "Monimatch" directional coupler,[1] which uses a capacitance-resistance bridge. The value of the terminating resistor, R_T, is critical and must be determined experimentally to obtain good bridge balance (null). The adjustment procedure and method of determining R_T in reference **1** should be followed before attempting to operate the instrument. *Editor.*

braid to pieces of stiff wire, which go to the connectors for the transmitter and transmission line. Next the diodes and resistor R_T are soldered to the ends and center, respectively, of the pickup inner conductor.

The pickup line and diodes were built in a shielded box, which connects in the transmission line between transmitter and antenna, or between transmitter and antenna tuner. The diodes I used were

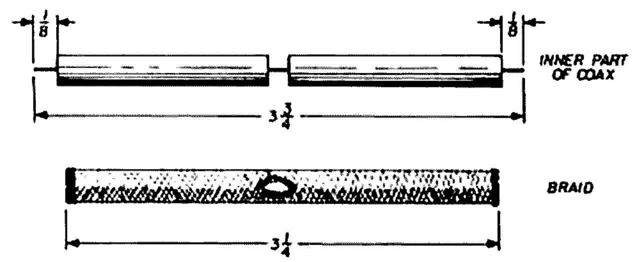


fig. 2. Construction of pickup line. Inner part of coax cable should be insulated where it protrudes through hole in braid; check with ohmmeter to make sure wire and braid are not short circuited.

Hoffman type 1N261, but the general-purpose 1N34A will do. A shielded two-conductor line was run to the indicator unit, which was built in an unshielded bakelite minibox. This box did not seem to pick up any stray rf. The shielded line need not be rf-type cable, since it carries only dc. The entire unit could be built in one box to eliminate a separate box and cable.

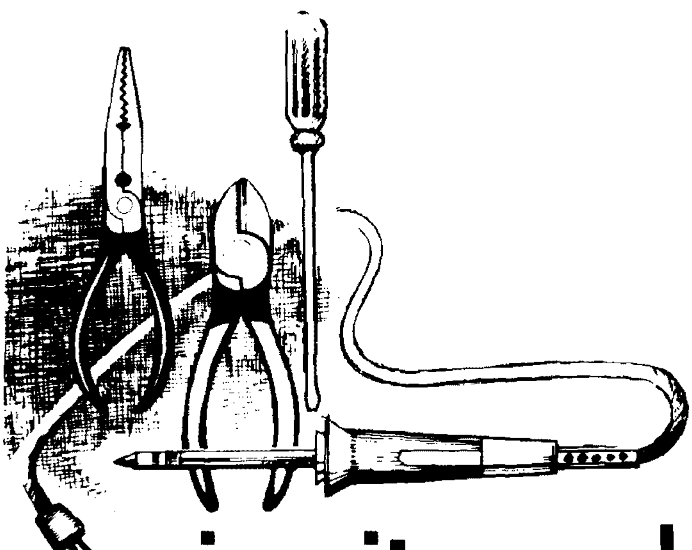
An ordinary 0-50 dc microammeter was tried and worked in the combined circuit without the switch. This would eliminate both the switch and a zero-center meter.

The bridge seems to have adequate sensitivity. Construction is straightforward and uncomplicated. Best of all, using it is as simple as tuning for maximum on the meter.

references

1. "The ARRL Antenna Book," 9th Edition, p. 135.
2. McCoy, "Limited-Space Antennas and Methods of Coupling," *QST,* February, 1969, pp. 24-27.

ham radio



circuits and techniques

ed noll, W3FQJ

experiments with phase-locked loops

Do you like to experiment? To improvise? You can start now with these little magic boxes, phase-locked-loop integrated circuits. I covered the fundamentals in last month's column; now you can get involved with some practical circuits.

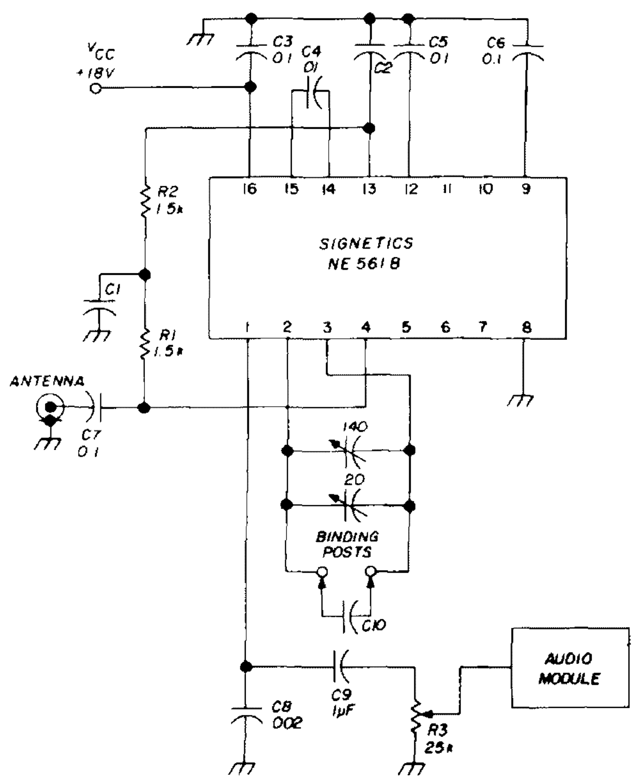
no tuned circuit

Can you conceive of a tunable receiver without a single resonant circuit. Take a look at **fig. 1.** All of the signals are fed into the magic box in one lump. The signal that is demodulated is the one with its carrier phase-locked to the voltage-controlled oscillator of the integrated circuit. You pick the station you wish to receive by changing the frequency of the oscillator. Even this oscillator does not have a resonant circuit because its frequency is determined by resistor-capacitor time constants.

The oscillator is stable if adequate carrier signal is applied to the phase-comparator, **fig. 2.** The output of the vco IC is supplied to a small solid-state audio module with an output rating of 1 watt; loudspeaker is a 4-inch pm unit. More than adequate volume level is obtained when using a good antenna system.

At this point in my experiments the unit has been used on the a-m broadcast band, the 160-, 80- and 40-meter amateur bands and the 31- and 49-meter shortwave broadcast bands. Vco tuning capacitor values and quadrature capacitor values are given in **fig. 1.**

Two variables, a 140 pF and a 20 pF, are mounted on the breadboard. The smaller capacitor is used for bandspread tuning. Two binding posts are included for convenience in changing the fixed-value capacitor associated with vco tuning. For operation on the broadcast band and 160 meters a 365- or 400-pF variable is appropriate. When using the values shown complete broadcast band



	C1	C2	C10
frequency	pF	pF	pF
BC band	150	150	200
			300
160 meters	100	100	150
80 meters	50	50	none
40 meters	50	50	none

fig. 1. Integrated-circuit phase-locked-loop receiver uses no resonant circuits.

operation is obtained by using several fixed capacitors, inserting them across the two binding posts when shifting among segments of the band.

A good antenna is important. A 160-meter dipole is excellent for 160 meters and the broadcast band; it also does well on 40 and 80 meters. However, dipoles cut for 40 and 80 meters do better on these bands as well as the shortwave broadcast bands.

If there is an overpowering local broadcast station that bulls its way through you may have to experiment with antenna lengths or install a resonant trap ahead of the antenna input.

more signal and selectivity

A simple resonant transformer at the input improves selectivity and blocks out any overpowering local, **fig. 3**. The low-impedance primary link matches a low-impedance antenna system. The secondary is resonant and is connected to the ic input by a low-value capacitor, C7. This capacitor prevents the low-impedance ic input from loading down the resonant circuit, insuring greater selectivity. A double-tuned input transformer would offer even greater selectivity.

When using a short random-length antenna a simple tuner such as the simple T-network in **fig.** 4 will deliver more signal. It is inserted between the receiver input and the antenna system.

More sensitivity. An amplifier ahead of the receiver (fet, bipolar or ic) will increase the sensitivity. However, the signal delivered to the PLL should not

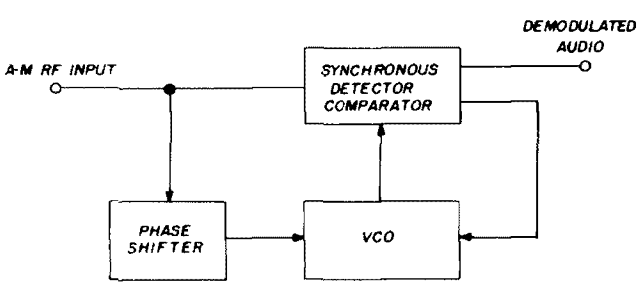


fig. 2. Basic phase-lock am demodulation scheme.

exceed 0.5 volts rms. Next month I will discuss my experiences with an amplifier.

Preset frequency operation. The PLL system lends itself to preset operation for particular stations on the broadcast band, WWV reception, net frequencies, etc. You need only use a switching arrangment and preadjusted trimmer capacitors as shown in **fig. 5.** You can then very quickly switch off the tunable position to one or more preset receive positions.

Idiosynchrosies. Everything is not ideal. Sufficient signal must be delivered to the input. Under weak signal conditions there is a swishing sound which results from the difference frequency between carrier and vco when there is an unstable lock.

There is also a hand-capacitance effect. This results from non-grounded capacitive tuning of the vco. Use a vernier dial and insulated shaft on the fine-tuning variable. The phase-locked loop ic also lends

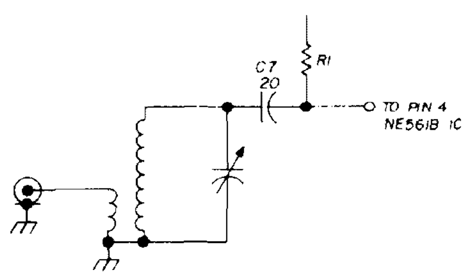


fig. 3. Resonant input circuit for improved selectivity.

itself to frequency change by voltage change, especially for fine-tuning. This may eliminate hand-capacitance effects. Suitable circuits will be discussed next month.

At higher frequencies input resonant circuits and tuning influence the vco frequency. It appears that the ideal arrangment would have an untuned amplifier just ahead of the ic input. Perhaps a better plan would be to use an isolating no-gain stage at this position and precede it with a tunable gain amplifier.

The PLL device in **figs. 1** and **3**, is suitable only for a-m demodulation, not cw or sideband. However, it can be used as a practical fm detector. A simple

switching arrangement would permit selection of either a-m or fm, making it ideal for 6- and 2-meter operation.

theory of operation

Important considerations in setting up the external circuit for the PLL synchronous a-m detector are the selection of the vco tuning capacitance and choosing appropriate values for the two RC combinations that establish the proper 90° phase relationship between the incoming carrier and the vco.

A very simple formula permits you to select a suitable value for the vco tuning capacitor:

vco capacitor (pF) = 300/frequency in MHz

Values should be calculated for minimum and maximum frequency you want to cover. For example, if the receiver is to tune between 1 and 2 MHz, the minimum and maximum capacitance values would be 150 pF (for 2 MHz) and 300 pF (for 1 MHz). A 125-pF fixed capacitor and

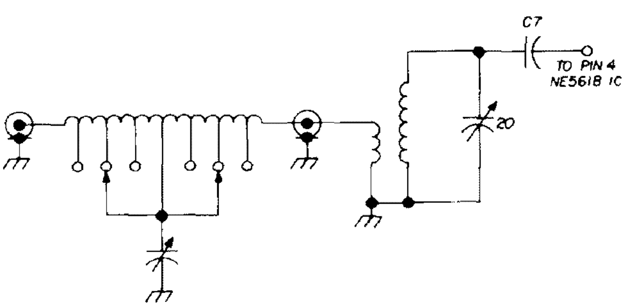


fig. 4. T-network antenna tuner for random-length antennas.

200-pF variable would provide band coverage plus a little bit of overlap at each end.

A net 90° phase shift is obtained by connecting two RC combinations in cascade. Each contributes a 45° shift. The phase shift of 45° is obtained at the frequency where the resistance and reactance are equal.

$$RC = \frac{1}{2\pi f}$$

In the actual calculation the frequency, f, is selected as the median frequency between the two desired frequency extremes. In our example this

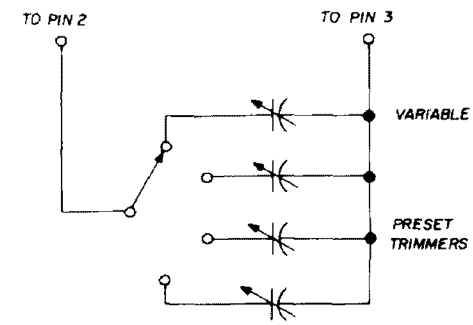


fig. 5. Presetting the phase-lock IC to specific frequencies.

frequency would be 1.414 MHz or:

$$f_O = f_H f_L = X\ 1 - 1.414$$

Where f_H and f_L are the high and low frequencies respectively.

If a resistor value of 2000 ohms is selected, the value of the associated capacitor must be:

$$C = \frac{1}{2\pi f_O} = 56\text{ pF}$$

Vco locking takes place over a considerable angular range; specifications are 90° ±30°. Therefore, it stays in lock between the high- and low-frequency tuning limits provided these limits are not spread too far from the median frequency. This means that the phase-shift of each RC combination should not be greater than ±15°. At these extremes the phase angle values would be 60° (45 + 15) and 30° (45 - 15). From a natural function table it can be seen that the tangents of 30° and 60° are 0.577 and 1.732 respectively. In practical applications this means that the reactance of the capacitor at the highest frequency should not be less than 0.577R, and the reactance at the lowest frequency should not be higher than 1.732R.

In our example, then, reactance at 2 MHz should not be less than 1154 ohms

(0.577 x 2000). At 1 MHz the reactance should not exceed 3465 ohms (1.732 x 2000). Calculations for 1 and 2 MHz show that 56 pF exhibits 2843 ohms reactance at 1 MHz and 1422 ohms at 2 MHz. These values are well within the angular locking requirements.

switching and linear amplification

A linear amplification system with efficiency as high as 90% has been developed by Brian Attwood of Mullard of England. This is indeed a startling figure when you consider the difficulty in obtaining 50% efficiency with conventional class-AB linear amplifiers.

The Attwood technique can be used with conventional amplitude-modulated rf signals but is especially adaptable to suppressed carrier and sideband modulation modes. Experimental work was done with solid-state devices but the idea is just as appropriate for vacuum-tube applications and perhaps even more so for hybrid combinations of semiconductors and vacuum tubes.

In the Attwood process the modulated signal is initially formed at low power levels in a conventional manner. The block diagram in fig. 6 shows the functional plan for a sideband transmitter. It includes the usual carrier generator and

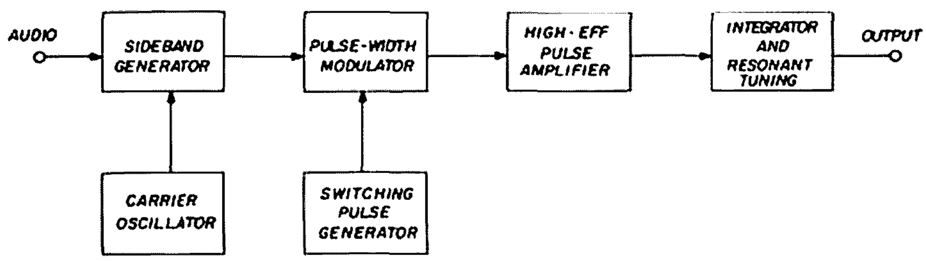
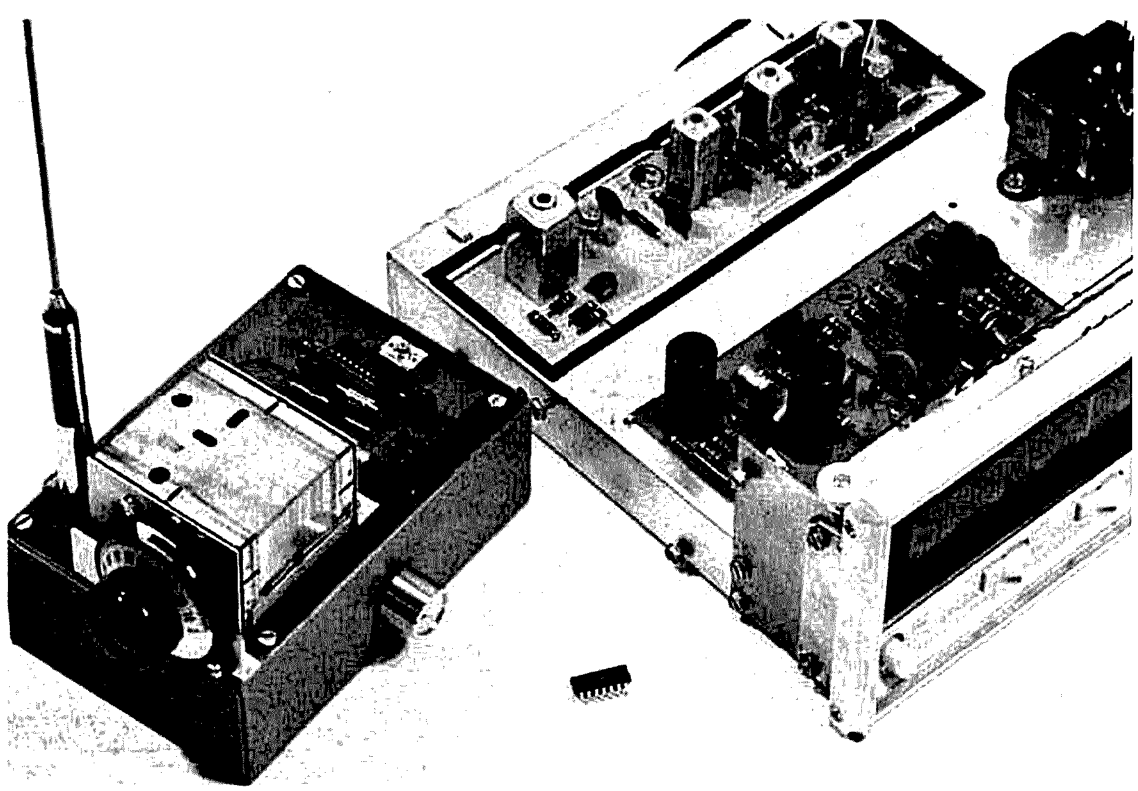


fig. 6. Basic arrangement for the Attwood high-efficiency switching modulator.



The phase-locked loop IC at the bottom center replaces the entire i-f/discriminator section outlined in black on the receiver chassis to the right. The unit at the left was built in the Signetics applications lab to demonstrate the reduction in size when a PLL is used in an fm tuner.

follow-up sideband modulator and mixer; the modulated signal at the transmit frequency is applied to a pulse-width modulator.

Two switching stages perform an additional modulation function. First there is a switching frequency generator which operates at a frequency a number of times higher than the transmit frequency. For good results this generator should be five times carrier frequency or higher although the system will function with the switching frequency only twice the carrier frequency. The switched modulation system produces an output that has the duration of its pulses varying with the modulating information.

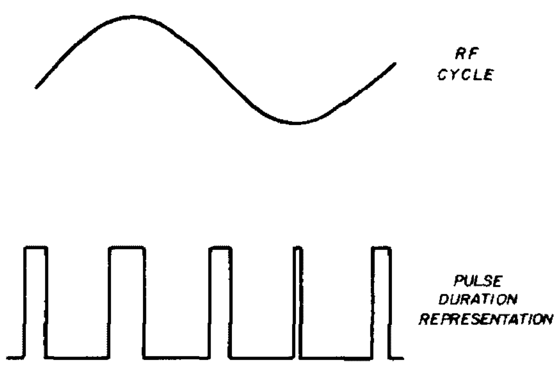


fig. 7. Example of pulse-width modulation.

A simplified drawing of a pulse-width modulation process is shown in **fig. 7.** As the modulated rf wave varies with modulation the pulse-width modulator generates a train of high-frequency pulses, with pulse width varying with modulation.

As an example, consider the single-tone modulation of a lower-sideband transmitter with a carrier frequency at 3.9 MHz. Under this condition a single rf wave with a frequency of 3.899 MHz (3.9 MHz minus 1000 Hz) is generated. As the 3.899 MHz rf wave goes through its cycle the pulse-width modulator produces a series of pulses. When the rf wave is on its positive crest the width of the output pulse is greater than the pulse that represents its negative trough. In fact, the width of the pulse varies in accordance with the instantaneous amplitude of the rf wave on each side of its zero axis. Of course, if there were voice modulation there would be complex pulse duration changes which would follow the amplitude gyrations of speech.

Note that pulse output has constant amplitude. All the desired information is in the form of pulse duration changes. The succeeding amplifier can then be made to operate at high efficiency because it can be designed to function as a *high-powered pulse amplifier;* pulse levels can swing between cut-off and saturation limits. The switched nature of the information conserves average power and results in greater power-handling capability and efficiency.

In an amateur transmitter it is conceivable that all stages prior to the final are solid-state; the final would be a high-powered vacuum-tube amplifier. This stage might use high-power television sweep tubes which are designed primarily for pulse and nonsinusoidal power amplification.

More power output could be obtained than is now possible using the same tube types as class-AB linears (common practice in many ham transmitters). The technique might lend itself to the design of mobile equipment with high output light-weight and minimum power demand.

After the pulses are amplified they are not transmitted; the signal is converted back to the form it had before it was introduced to the pulse-width modulator. This can be done with a suitable resonant system since resonant circuits have energy-storing ability and function as effective integrators. Thus, the pulse information is stretched out and converted to sinusoidal form.

The output circuit, of course, must be designed to remove and attenuate any switching frequency components which are on a higher frequency than the transmit frequency. It is conceivable that a suitable multisection pi-network output system would do a satisfactory job. Perhaps an m-derived addition may be necessary to thoroughly notch out the switching frequency.

Because of the requirement for the high switching frequency this modulation mode is currently most suitable for low-frequency operation. However, vhf possibilities exist when you consider that low-cost ics are now available with switching rates up to 500 MHz or so.

version I used one of the commonly available 1-watt output audio modules. A transistor audio-output transformer connected in reverse (low impedance to high impedance) permitted the necessary voltage drive and reasonable match to the source input circuit; the center tap of the

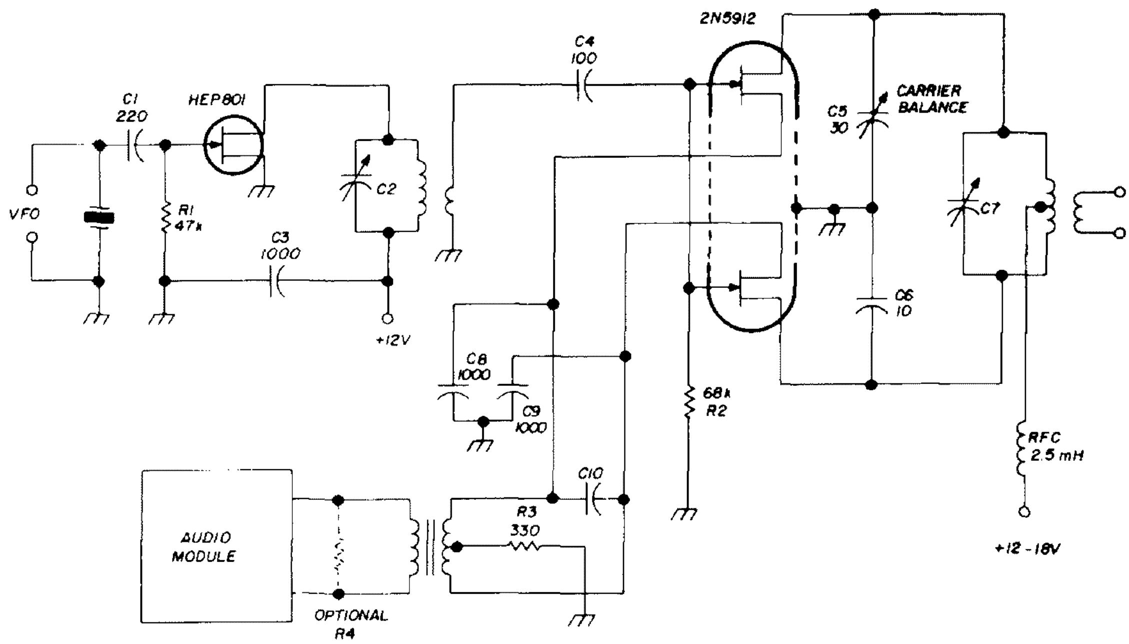


fig. 8. Dual-fet balanced modulator.

dual-fet balanced modulators

A dual field-effect transistor consists of two fets with identical characteristics mounted in the same case. Such units are designed mainly for use in differential amplifiers, but they are also ideal for balanced modulator and demodulator circuits. Devices are available which function up to several hundred megahertz.

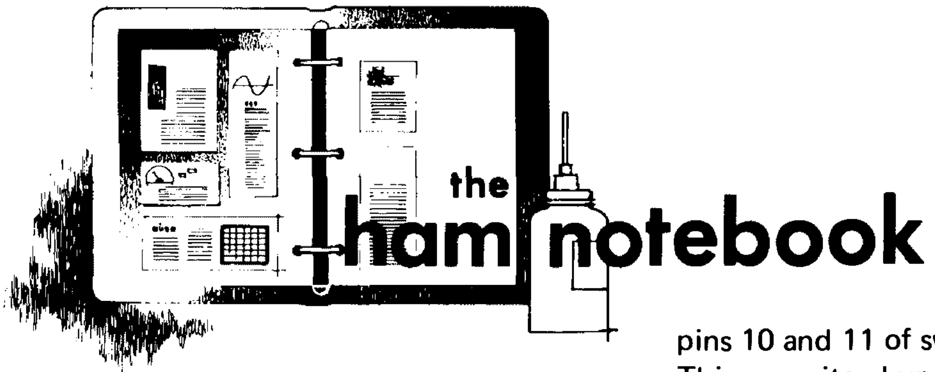
An effective circuit that I have operated on 80 through 10 meters using plug-in coils is shown in **fig. 8.** In this circuit the paralleled gates of the Siliconix 2N5912 dual fet are supplied with carrier from a crystal oscillator. The drains are connected in push-pull to obtain carrier cancellation. Removal of the crystal from the oscillator permits use of a stable vfo. Output level is correct for driving a low-power vacuum-tube linear amplifier using a 6AK5, 6BA6, etc.

The modulating signal is applied to a push-pull source circuit. In my pegboard output winding permitted a balanced feed system. A capacitor across the secondary (C10) is used to control the high-frequency audio roll-off.

The average audio module may supply more audio than is necessary. If this is the case a loading resistor (R4) of 12-ohms or higher (depending upon how much the audio amplitude must be cut back) can be connected directly across the output of the module. The microphone gain control is connected between the mike terminals and the audio module input.

Carrier balance is handled by capacitors C5 and C6. The trimmer capacitor is adjusted for minimum carrier output; carrier suppression is excellent. No carrier reduction control was included in the source circuit although some additional suppression may be possible. However, if the dual fets are really identical it should not be necessary.

ham radio



32S-3 audio

The two audio amplifier stages in the Collins 32S-3 exciter run full-blast, inasmuch as the *mic gain* control follows them. With high microphone levels the second audio stage is overdriven and adds distortion; this can be verified by placing a voltmeter between V1B pin 6 and ground. This distortion may be reduced by allowing slightly higher *mic gain* settings, and adding desirable degenerative feedback in the audio stages to reduce the level at which the second audio stage operates.

Collins Service Bulletin No. 2 (dated 21 September 1967) says, "disconnect and discard the bus wire connected between tube socket XV1, pin 8, and switch S8C, pin 2. Replace the bus wire with a new 680-ohm resistor, R1." R1 is ½ watt. This change was incorporated in the fifth edition of the 32S-3 transmitter instruction book.

Incidentally, Collins Service Bulletin No. 1 was revised on 13 October 1967, making a number of changes. After tests, I restored the 1-megohm resistor R101 in the first audio stage, V1A, rather than use the 100k resistor. Many of the changes already were incorporated in my equipment, which is covered by the third edition of the instruction book (dated 15 September 1964).

My attention was called to another change to increase degenerative feedback. The third edition of the instruction book shows a 20-μF capacitor, C183, between pins 10 and 11 of switch S9F and ground. This capacitor bypasses the second audio cathode resistor, R4. It did not appear in the fifth edition of the book; Collins recommended that it be disconnected. The capacitor is located conveniently behind the words *freq control* on the upper left of the front panel so the change can be made without removing the set.

If there still is excessive audio gain Collins has recommended a ½- or a 1-megohm resistor between pins 1 and 6 of the audio stages, V1. This is between the two plates. I tried it using a Vector test socket under the tube, and found that it is desirable, resulting in excellent audio quality after removing the capacitor mentioned above. The level in the second audio stage was reduced to a distortionless level and the *mic gain* was at a reasonable half-scale position. I had been using the Electro Voice 676 dynamic cardoid microphone, with 10-dB roll off.

Bill Conklin, K6KA

narrow-shift RTTY reception with Heath SB receivers

To receive 170-Hz narrow-shift RTTY signals with the Heathkit SB-100, SB-101, SB-300 or SB-301 using the optional 400-Hz filter, the bfo crystal must be 3393.19 kHz. In the SB-series receivers the lower-sideband crystal is 3393.6 kHz; the frequency may be lowered to

3393.19 kHz by placing a small capacitor from the grid of the oscillator tube to ground. This shunt capacitor must be chosen so that both the *mark* and *space* tones have equal amplitude; 20 pF worked satisfactorily in several sets tried here.

In the SB-300 and SB-301 receivers the 3393.6-kHz crystal must be placed in the upper-sideband crystal socket so it will be selected when the 400-Hz filter is switched in.

Robert Clark, K9HVW

dynamic transistor tester

If you already have an oscilloscope on your workbench here is a simple dynamic transistor tester which will measure both in- and out-of-circuit transistors. Total cost of the device is less than $5. In addition, this tester will check diodes, although it cannot be used with mosfets, junction fets or uhf varistors and diodes.

In the circuit in **fig. 1** a small ac voltage is fed to the transistor junction. This alternately forward and reverse biases the junction. With the oscilloscope test leads open-circuited a horizontal trace is displayed; when the test leads are shorted a verical trace is shown. A good transistor junction, either emitter-to-base or base-to-collector, when connected between the two test leads, will show a sharp-cornered trace as shown in **fig. 2D.** A rounded corner (**fig. 2E**) indicates

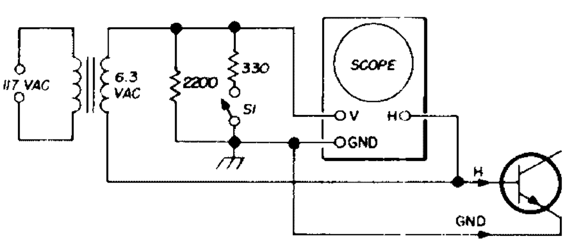


fig. 1. Simple dynamic transistor checker tests both in- and out-of-circuit transistors and diodes.

leakage current. Junction resistance shows up as a sloping vertical trace (**fig. 2F**).

To calibrate the oscilloscope, switch the horizontal sweep to *external,* plug in the test leads, and with the leads open-circuited, adjust the horizontal control

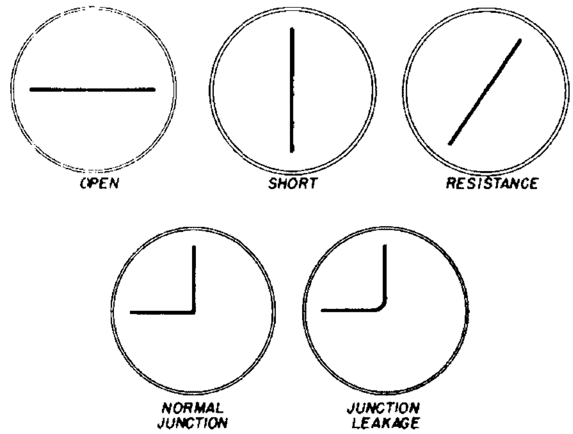


fig. 2. Basic traces obtained with the test circuit of fig. 1.

for approximately 2/3 scale deflection. Now short-circuit the test leads and adjust the vertical control for approximately 2/3 scale deflection. The equipment is now ready for testing.

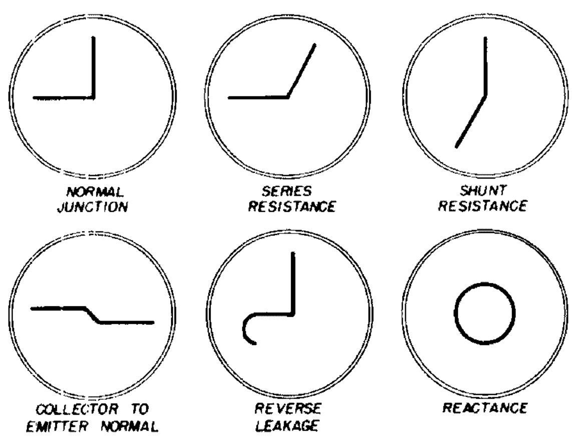


fig. 3. Typical circuit traces provided by the dynamic transistor tester.

Power transistors require more test current — this is provided by closing switch S1. Most in-circuit transistors will indicate some resistance such as shown in **fig. 3B** or **3C.** This transistor tester will not check an in-circuit transistor when the impedance across the junction is extremely low.

Vern Epp, VE7ABK

radio-controlled morse sounder

The circuit shown in fig. 4 provides operation of a Morse telegraph sounder off the air. There are many amateurs who are also Morse operators and would like to hear the music of a telegraph sounder again. With this circuit the relay picks up with the reception of a signal.

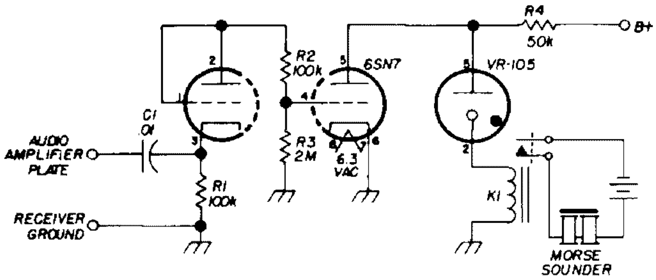


fig. 4. Circuit for operating Morse sounder off the air. K1 is 10,000-ohm spst relay.

The first half of the 6SN7 is connected as a diode rectifier. The incoming audio signal is rectified to bias off the second half of the 6SN7, which is a clamp tube. When the tube is clamped off, current flow stops, and the voltage at pin 5 of the VR105 increases, igniting the VR tube and causing the relay to close. If more audio gain is desired a one-stage amplifier could be inserted between the receiver and the first half of the 6SN7 although this should not usually be necessary.

For proper operation, current flow through the relay should be measured between the relay coil and ground. The value of R4 may have to be reduced to approximately 40k for snappier sounder operation. Capacitor C1 may be increased to 0.1 μF if desired. Total current drain from a 250 to 300-volt receiver power supply is about 50 mA.

Jack Proefrock, K6QEQ

nylon guy rope

Most amateurs who have experimented with antennas realize the problems generated when steel guy wires are used in close proximity to the antenna they support. Swr and pattern distortions are just the beginning.

After a recent residence move I decided that the only location on the property suitable for a tower was the roof. This meant that guying would be essential. In an effort to bypass the usual interference problems I considered two possible remedies. One was the use of strain insulators to break up the guy wires into non-resonant lengths. I ruled this out for several reasons: time involved, added potential weak links and the increased cost.

The second consideration was the use of nylon rope. I finally decided on a rope with a working strength of 148 pounds (in excess of 750 pounds test). My

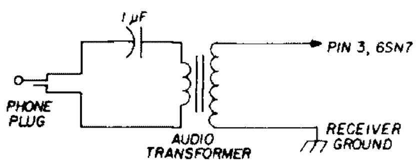


fig. 5. Interface circuit for plugging Morse sounder circuit into your receiver.

24-foot aluminum tower went up from its tilt-over base without a hitch. However, when it was time to begin tightening the nylon guy ropes, the first hint of trouble appeared. We couldn't tighten each leg enough; the nylon, especially the end knots, was stretching.

With hopes of rectifying the situation, we switched from knotted ends to split bolts. This helped, but not as much as I would have liked. Even so, I decided to give it a trial run. I kept the installation *status quo* for the entire winter and found that nylon was every bit as strong as first anticipated, and held in winds gusting over 90 mph. The nylon guys showed little, if any, signs of weathering; the split bolted ends were still secure.

After the tower was up a short time I discovered that strength was not the only criteria to consider. Although the tower guys held safely in high winds, stretching became more and more evident. The tower and array often swayed violently as

if suspended by six huge rubber bands. This generated a lot of unpleasant noise and may have weakened the base mount. Needless to say, this sort of installation was unacceptible.

As is often the case, the eventual cure was a compromise. Nylon rope was used for the top set of guys, while steel was chosen for the lower supports. This arrangement reduces noise and provides acceptable strength with minimum guy-to-antenna interaction.

Morrie S. Goldman, WA9RAQ

drill guide

While building various projects I occasionally find it necessary to accurately drill a hole in the center of a ¼-inch volume-control shaft. I first obtain a short length of soft-iron stock, 1-inch square or smaller, depending on what is available; then I drill a hole through the iron with a small drill (7/64-inch or 6-32 tap drill). This can be done with a hand drill.

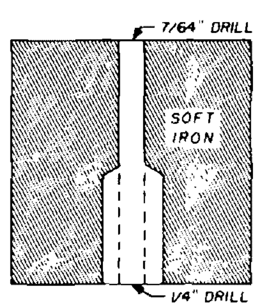


fig. 6. Simple jig for drilling hole in the center of a ¼" shaft.

Using this small hole as a pilot hole, I drill halfway through the metal with a ¼-inch drill. This completes the drilling jig. To use the jig, insert the ¼-inch shaft into the larger hole and drill through the smaller. Centering accuracy is quite good with this technique and it does not require any elaborate tools.

This method is suitable for drilling other size shafts as well. Simply choose the correct size drills for each of the holes.

Felix W. Mullings, W5BVF

improving linear amplifier performance

In a recent article[1] it was noted that unsatisfactory performance of transistorized equipment could often be traced to poorly regulated power supplies. This is also true of linear amplifiers. The point was very much stressed when ssb first became popular but is often overlooked by the new generation of amateurs. My experiences in performing a simple test for linearity and observing the increase in power output that resulted from improving power supply regulation may be instructive.

My linear, the Heath HA-10, had a single 8-μF capacitor in the power supply. Rf output is normally monitored with a Heath HO-10 Monitor Scope. When I keyed the exciter with a succession of dots at about 40 wpm it was obvious from the scope display that there was an enormous variation in output within a single dot, following an approximately sinusoidal pattern. It illustrated precisely the conditions described years ago in *GE Ham News*.[2]

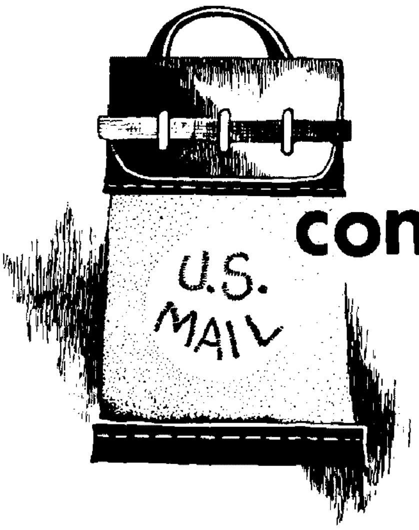
The correction was simply to add 20 μF of filtering capacitance (there is plenty of room in the spacious old HA-10 linear). The result was that both long and short keyed pulses were nearly flat on the monitor scope. Furthermore, there was a substantial increase in average power output. I suspect that improving the dynamic performance of a linear amplifier's power supply will be an economic way of increasing power for many amateurs.

references

1. W. Wildenhein, W8YFB, "Regenerative Detectors and a Wideband Amplifier for Experimenters," *ham radio*, March, 1970, page 66.
2. "About Power Supplies: Better Dynamic Characteristics Mean Better Output on CW, AM and SSB," *GE Ham News*, January/February, 1954.*

Guy Black, W4PSJ

*Copies of this article are available from the author for 10 cents and a self-addressed stamped legal-size envelope. *editor.*



comments

collinear antenna

Dear HR:

I doubt that WB6KGF is getting 14-dB gain from the 4-element 144-MHz colinear described in the May, 1971 issue. A single 5/8-wave section gives 3-dB gain over a ground plane; theoretically, two sections provide 6-dB gain, and four sections yield 9 dB – *if* current is equal in all sections. When fed at the bottom, top sections have less current. I would guess that 7-dB gain would be a more realistic figure.

William I. Orr, W6SAI
San Carlos, California

Dear HR:

W6SAI is quite right in saying that it is impossible for the gain of a 4-element colinear to be entirely responsible for the 14-dB increase in field strength. As I said, some of this increase could be attributed to an increase in transmitter power input; I should have emphasized this more.

During the tests the colinear was at the end of a long run of old RG-8/U coax (over 100 feet) with fixed 50-ohm transmitter output. However, I don't think gain would be as low as 7 dB – don't forget, mutual coupling between elements plus the low-loss characteristics of the copper-tubing elements would reduce any current imbalance caused by feeding the antenna at the end as opposed to the center.

The antenna system is highly resonant with circulating currents far exceeding the input rf current. Unfortunately, since the antenna described in the article is in Viet Nam, it is not available for further tests with input power held constant.

Robert A. Dahlquist, WB6KGF
Kingsburg, California

old-time radio

Dear HR:

K4NW's excellent article, "Those Were the Days," brought back memories of the good times I used to have with ham radio. I have had a transmitter of some sort on the air since May, 1913, starting out with a Ford spark coil powered by Columbia 6 dry cells, obtained from the telephone company when they installed new ones. I will never forget the thrill that came with my first contact – it was only seven miles away, but what a thrill!

My first receiving crystal was homemade. I used a teaspoon for mounting; the holder was a piece of pure lead about the size of a large pea. The crystal itself was made by melting flower of sulphur until it formed a crystal. Believe it or not,

it worked very well at the time.

In the early 1920s I built one-, three- and five-tube receiving sets, as well as a visual scope. For the scope I used an old Baldwin speaker unit, mounting it on the bottom of a tin can with lead weights. A toy balloon was stretched across the top of the can; a teaspoon of mercury was put in the center of the balloon. With a light bulb shining on the mercury, and a mirror arranged so the reflection from the mercury pool was projected onto a ground-glass screen, I could obtain good pictures of waveforms.

Charles M. A. Shade, W5NKA
Giddings, Texas

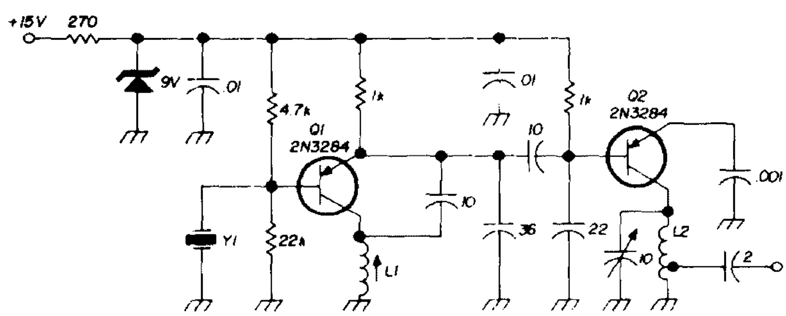
two-meter converter

Dear HR:

I want to thank you for the WB2EGZ article in the February issue of *ham radio.* I have done a lot of playing with solid-state converters over the past several meters around this part of the country than anything since the launch of OSCAR V. I have not had a chance to measure the noise figure accurately, but the five converters I have checked exhibited noise figures from 2.5 to 3.0 dB.

Although I have never had much luck with printed circuits on two meters the WB2EGZ design works very well. I did have to add a shield between the input and output of each of the rf stages; this cured the self oscillation.

I also modified the oscillator circuit (**fig. 1**); a number of amateurs wanted to use their existing converter crystals to keep construction cost down. This circuit is simple to duplicate, is very stable, and has a lot of output as a frequency tripler. You will notice that there is no resistor in the emitter circuit of Q2 — only a .001 capacitor. The emitter junction is used as a varactor, a trick used in some of the Parks converters. I have used this basic oscillator circuit for several years because it is simple and easy to build.



i-f	14 MHz	28 MHz	30 MHz
Y1	43.333 MHz	38.666 MHz	37.833 MHz
L1	11 turns no. 28 on 3/16" form	12 turns no. 28 on 3/16" form	12 turns no. 28 on 3/16" form
L2	6 turns no. 16, 1-turn spacing, airwound, 1/4" ID		

years, and I had what I thought were good designs. They worked well, provided good gain, and had good noise figures, usually around 2.5 dB.

I have now built several converters using the WB2EGZ design and they all perform very well. The converters were easy to duplicate, which speaks well for the basic design. In fact, the WB2EGZ article has created more talk on two

The transistor lineup in my converter has an RCA 40673 at the input, a 3N140 in the second stage and a 3N141 mixer. The transistors in the oscillator and tripler can be any good pnp devices. I have used a number of different transistors including the 2N3284, TIX10, and M124; nearly anything seems to work.

John C. Fox, WØLER
Minneapolis, Minnesota

rf interference

Dear HR:

When selecting the transistor radio as an i-f for the converter used as a noise locator (*QST,* June, 1966), it will prevent birdies if a trf instead of a superheterodyne receiver is chosen. However, be sure that the trf set has an unused broadcast frequency available on it without interference.

The *QST* converter has no connection between the oscillator and the mixer in the high-frequency range. Tests with a signal generator resulted in adding a coupling capacitor for some bands in order to obtain adequate sensitivity. Also, the whip antenna was too small. A bamboo pole, wound loosely with wire, with a long lead to a plug, provided sensitivity comparable to the home receiver. In a few cases, a wire coil used as a loop antenna, even on the fm band, could be used for determining the direction of the noise.

Bill Nelson, WA6FQG, emphasizes the need to shift to a higher frequency as the noise is approached, thus limiting its area. On the 7- and 3.5-MHz bands, the noise may be heard for a mile or two, but this can happen even on 28 MHz if the noise source is high and in line of sight. As the receiving frequency is increased when the noise is louder, there may be a substantial reduction in noise range, permitting accurate noise location. Some noises, however, may be heard in a relatively narrow frequency range. Fluorescent lights are likely to have a high-frequency cutoff around 7.5 to 8 MHz. It is helpful to have a general-coverage noise locator because of the possible frequency limitations of the noise.

On the upper high-frequency bands there may be standing waves so it is necessary to walk beyond the point of minimum noise to determine whether there is another peak further along – possibly a stronger one! Reduction of the rf gain may help particularly where the agc cannot be disabled. Beware of increased noise levels near power poles having a ground wire; this just adds antenna length to the noise-locator receiver.

It is well to have a sledge-hammer, or at least a large rock, with which to "thump" a power pole in the noisy area. Usually, the pole "thumped" will cause the noise to break up (and sometimes to come on, or go off), but movement sometimes travels down the wires to an adjacent troublesome pole. Thick poles don't respond to this treatment, but frequently it is possible to make a guy wire swing sufficiently to produce the necessary movement of a pole to affect the radio interference.

New electric lines appear to be worse than old ones. Here, this has resulted from installation where the nuts were left loose all over a pole! Sometimes, a noisy pole again becomes noisy a half-year later. One capacitor bank on a pole near my home has created noise three different times.

One swimming-pool pump motor put out noise for a hundred feet. It was reported to the owner. Within about two weeks, it failed completely and was replaced.

The Navy purchased shield boxes for fluorescent lights; in addition to line filters and metal shields, the light passed through a conducting-glass cover. They were very effective, even in a radio receiving location underground.

Bill Conklin, K6KA
La Canada, California

vhf fm receiver

Dear HR:

In the September issue there was a 2-meter fm receiver project that really took my fancy. I built it, and it worked so well I wanted to tell you how pleased I am with the finished receiver.

Since I used printed-circuit boards I only had to worry about collecting the necessary parts, and alignment. Collecting parts was the hardest part of the whole project – I enlisted the help of everyone I know. Most of the parts were not too difficult, but the Sprague integrated circuit and i-f cans were tough. I also had

some difficulty finding shields for the handwound coils. I finally spent a dollar and ordered six inexpensive i-f cans from a dealer on the East coast; after discarding the goodies inside, the cans worked fine for coil shields.

The i-f transformers specified in the original article were too expensive for me. The author now advises the use of Calectro D1823 i-f transformers. These work fine and cost less than one dollar each. The Sprague integrated circuit was the most elusive rascal I have ever hunted, but I finally found one.

The sensitivity of the receiver really surprised me – 0.3 μV for full quieting. If you decide to build this receiver I hope you have as much fun as I did. It is a very worthwhile project, and one that will be in constant use.

Larry Pepple, WA9RRZ
Fort Wayne, Indiana

audio filters

Dear HR:

That simple audio filter described by W4NVK in the October issue of *ham radio* really works. I put one together out of the junk box, about 0.3 μF across the 88-mH toroid, and it entirely cuts out the QRN. Tuning was very sharp – I estimate about 100 Hz wide; so sharp, in fact, that I had to tune with the receiver pitch control. There is about 10 dB reduction in volume, which is not too bad.

I put it together with no solder, just jumpers, and it about floored me when I kicked it into the circuit. I worked a novice on 80 meters with no problem – when I turned the filter off he disappeared into the QRN.

I don't get excited very easily, but this filter really works. My SX-101A and 500-Hz filter with bridged-T notch are sick next to it. I added a 1000-ohm pot across the filter to adjust the amount of filtering – at zero resistance the filter is shunted out, with maximum resistance you get full filtering.

Duane Schnur, WB8EEJ
Caro, Michigan 48723

1296-MHz moonbounce

Dear HR:

I am using two Motorola 1N5150 varactor diodes to produce 10 watts output on 1296 MHz; this will be used to drive a pair of 3CX100A5s to 100 watts into a circularly symmetric feed on my 28-foot dish. Hopefully I will be able to work G3LTF and W2NFA via moonbounce during March.

Peter, G3LTF, has sent me a preamp which has a measured noise figure of 5.3 dB and has been used to receive his own echoes on his 15-foot dish. The preamp will be mounted at the feed point with the converter in the shack. I have some 1-5/8" Andrews heliax which will be used on the transmitting side, and as my dish is more accurately constructed than Peter's I should be able to hear him between 6 and 10 dB above the noise. My own echoes should be 2 to 3 dB weaker as he is running four 3CX100A5s and inferior cable.

I hope that some modifications to my dish will be made and the dish remounted by late February or early March. The polar mount will enable accurate tracking automatically any time the moon is above the horizon.

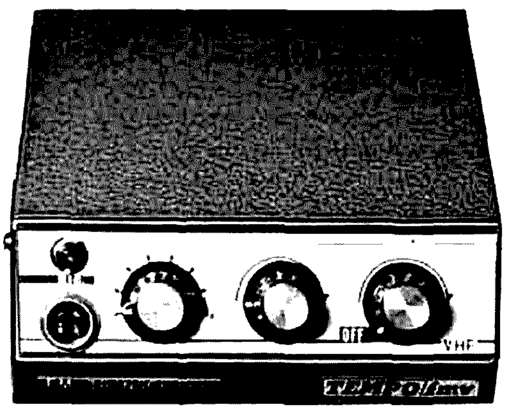
I also have plans for 32 18-foot long crossed *Swan* yagis mounted 15-feet apart in both planes and phased for vertical, horizontal, clockwise or counter-clockwise circular polarization. This will be polar mounted on another tower that I hope to have operational by the end of this year.

My ultimate aim is to provide moonbounce antenna capability on 144, 432 and 1296 MHz to any interested vhf group or individuals in Australia at no charge to foster international goodwill and promote moonbouncing to many who would otherwise be unable to partake of this fascinating aspect of amateur radio.

Ray Naughton, VK3ATN
Birchip, Australia

new products

vhf fm transceiver



The new Tempo/fmv from Henry Radio is setting new highs in performance and value. The solid-state two-meter transmitter features 12-watts rf output with spurious signals more than 60-dB down. Frequency deviation is adjustable from 5 to 15 kHz. The dual-conversion receiver has 0.6 μV sensitivity for 20-dB quieting (0.3 μV usable threshold). Selectivity at full quieting is ±6 kHz at -6 dB and ±15 kHz at 170 dB. Receiver audio power output is 1 watt.

The Tempo/fmv provides eight-channel coverage. Power supply requirements are 12 to 15 Vdc at 2 amps; the unit weighs 4.5 pounds. Extra features of this fm transceiver include a build-in metering test socket, an operation/maintenance manual and an optional test-set accessory. The built-in test socket can be used with a sensitive microammeter to monitor all stages including the discriminator. The large instruction manual covers complete checkout and alignment of the transceiver. The optional test set mates with the built-in test socket and includes a sensitive microammeter.

The Tempo/fmv two-meter fm transceiver is priced at $249 including microphone from Henry Radio, 11240 West Olympic Boulevard, Los Angeles, California 90064. The optional test-set accessory is priced at $29. For more information use *check-off* on page 126.

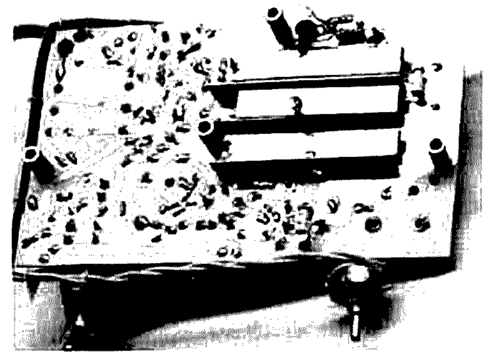
gem-quad antennas

The fiberglass quad antennas from Gem-Quad in Canada offer a number of interesting design features, including light weight, a cone-shaped design that maintains critical measurements under severe weather conditions, non-corrosive construction, single 52-ohm feedline on all bands, low swr, and nylon tension tubes at the corners of the quad to eliminate sharp angles in the wire, thus assuring longer antenna life. Gem-Quads come in two-, three- and four-element arrangements for use on 10, 15 and 20 meters. Forward gain for the 2-element quad is reported to be 8 dB on DX signals; with an optional third element (which may be easily installed with no conversion) gain is increased to about 8.9 dB. Front-to-back ratio of the 2-element model is 25 dB; for 3 elements, front-to-back is 30 dB.

The Gem-Quad antenna is designed for optimum performance and can easily be rotated by an ordinary tv rotator. When properly assembled, the Gem-Quad is capable of withstanding winds up to 100 mph. The 2-element quad is $107.00 complete; 3-element quad, $167.00; and 4-element quad, $227.00 complete. A third of fourth element, if purchased separately, is priced at $60.00. For more information write to Structural Glass Limited, 20 Burnett Avenue, Winnipeg 16, Manitoba, Canada, or use *check-off* on page 126.

More Details? CHECK–OFF Page 126

432-MHz converter



The new Janel 432-MHz converter provides many design features for the 432-MHz operator, including noise figure of 5.5 dB, any 4-MHz band from 420 to 470 MHz, 35-dB gain (adjustable) and image rejection (with 28-MHz i-f of 40 dB; i-f rejection is 95 dB. The new high performance converter uses all silicon and mosfet transistors, high-Q air-line tuned circuits for low loss and built-in zener-regulated power supply for extreme frequency stability; the built-in power supply may be used for 117 Vac or 12 Vdc operation.

The converter circuit uses five npn bipolar silicon transistors plus one mosfet. The rf amplifier uses a 40235 in a common-emitter configuration with a broadband input circuit to tune out input reactance. Two silver-plated stripline output circuits provide maximum selectivity and image rejection. The local oscillator chain begins with a crystal in the 100-MHz region, thus reducing the number of multipliers and spurious responses.

A 40235 oscillator excites two 40235 doublers to obtain injection to the base of the 40235 grounded-emitter mixer. The output of the mixer is capacitively coupled to the three-stage mosfet i-f amplifier that provides zero-to 27-dB i-f gain by adjustment of a control on the front panel.

I-f output frequencies available off the shelf are 26-30 MHz, 28-32 MHz and 50-54 MHz. One-year guarantee. $64.95 from Janel Laboratories, Post Office Box 112, Succasunna, New Jersey 07876. For more information use *check-off* on page 126.

solid-state qrp projects

QRP operation is rapidly becoming one of more popular facets of amateur radio, largely because it means a return to homemade equipment and economical operation. This new book by Ed Noll, W3FQJ, covers a variety of solid-state transmitting gear, with power ratings from less than 100 milliwatts up to about 20 watts. A variety of solid-state crystal oscillators and vfos are included as well as multistage cw transmitters. There are also a-m and single-sideband circuits.

The emphasis in this book is on solid state with circuits using bipolar transistors, field-effect transistors and integrated circuits. Also included are an introduction to practical solid-state circuit theory, QRP test gear and simple antennas for QRP operation.

128 pages, softbound. Published by Howard W. Sams & Co., Inc. $4.25 from Comtec Books, Box 592, Amherst, New Hampshire 03048.

RTTY control terminal

The new WCI phase-locked-loop detector-type RTTY demodulator converts the audio-frequency shift tones from a communications receiver to dc pulse data that operates a teleprinter. This demodulator has the unusual ability to track the input audio signal frequencies automatically if they change frequency due to transmitter or receiver drift. In addition, this unit will automatically copy any shift from 100 to 1000 Hz. Also included in the detection circuitry is an automatic threshold computer.

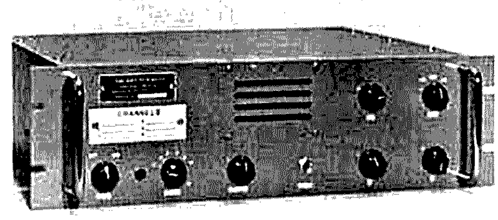
If the input signal is degraded past a preset level (due to a fading or a very poor signal-to-noise ratio) an automatic noise squelch circuit places the unit in *mark hold* to hold the selector-magnet armature closed to prevent the machine from printing unwanted erroneous characters.

The RCT-2D RTTY control terminal uses all solid-state circuitry with plug-in printed-circuit boards. Each printed-cir-

cuit board has built-in top-access test points and adjustments for easy maintenance. The RCT-2D can also be supplied with digital autostart and selective call. The digital autostart keeps the teleprinter motor turned off until a message is being received. Within the first three characters the motor turns on and the message is printed. Ten seconds after the last character is received, the motor is turned off again.

The RCT-2D RTTY control terminal is available from WCI, Post Office Box 17, Schaumburg, Illinois 60172. For those amateurs who want to build the WCI phase-locked demodulator themselves, the company is offering a kit of pre-drilled circuit boards with instructions for $45. For more information use *check-off* on page 126.

high-frequency receiver



The new Galaxy FFR-230/6 receiver is completely solid state and crystal controlled for operation on any six channels in the 2- to 18-MHz range. Front panel controls are provided for channel selection, mode—usb and lsb (2.1 kHz bandwidth) and cw (1.5 kHz bandwidth), clarifier (ssb), bfo, af and rf Gain. Features include rack mounting, internal speaker and 117-Vac power supply, external muting and audio (4 ohms and 600 ohms), modular construction and plug-in circuit boards for changing frequency. Priced at $950.00. Optional filter bandwidths are available.

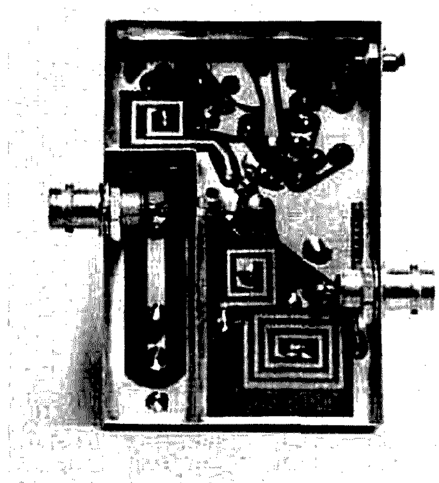
For more information on the new Galaxy FFR-230/6 receiver, use *check-off* on page 126 or write to Galaxy Electronics, Subsidiary of Hy-Gain Electronics Corporation, Route 3, Lincoln, Nebraska 68505.

More Details? CHECK—OFF Page 126

144/432-MHz transverter

The new Braun TTV-1270 solid-state transverter seems like the ideal way to become operational on 432 MHz. The TTV-1270 was developed specifically for mobile, portable and field-day operation, although it is also suitable for low-power fixed-station operation on 432 MHz.

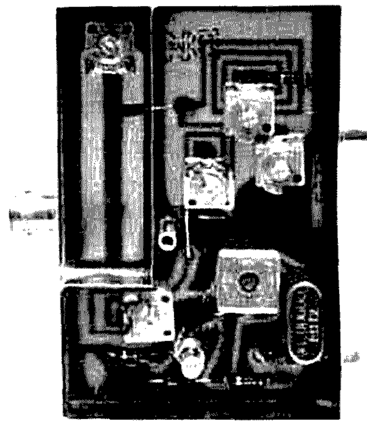
To put the TTV-1270 on the air simply connect it between a 2-meter a-m, fm or cw transceiver and a 432-MHz antenna. No antenna switching is required; a 12-volt dc power supply is needed only for receive. Input frequency range is 144 to 146 MHz; output range is 432 to 438 MHz.



The transmit section of the TTV-1270 transverter uses a factory-aligned varactor tripler. Input power should be from 50 mW to 1.2 watts. Output is 30 to 720 mW at 60% efficiency. In the receive mode the output of a 96-MHz voltage-regulated crystal-controlled oscillator is tripled to 288 MHz. This signal serves as the injection for a 432-MHz diode mixer stage. Noise figure on receive is 9 dB; receiver gain is -9 dB. To change from transmit to receive, the 12-V power supply is connected; current drain in the receive mode is 12 mA.

The Braun TTV-1270 is built into a silver-plated enclosure with rf-tight partitions. The stripline 432-MHz circuits are

silver-plated on glass-epoxy printed-circuit board. BNC connectors are used for both input and output connects. The transverter is factory-aligned for 50-ohm input and output impedance. Slight improvement in performance can sometimes



be obtained by retuning the unit with your two-meter source and 432-MHz antenna; complete instructions are provided with the unit.

The Braun TTV-1270 is specially introductory priced at $37.50 from Spectrum International, Post Office Box 87, Topsfield, Massachusetts 01983. For more information on Braun vhf and uhf equipment, use *check-off* on page 126.

transistor substitution handbook

Although bipolar transistors are noted for their low failure rate occasionally they do have to be replaced. As long as the specific type number required is available replacement of the transistor is no problem because a duplicate of the original should be used whenever possible. All too often, however, an exact replacement cannot be obtained without considerable delay. Furthermore, the great variety of transistor types make it difficult to determine which transistor can be substituted for the original.

The new 11th edition of "Transistor Substitution Handbook" is possible be-

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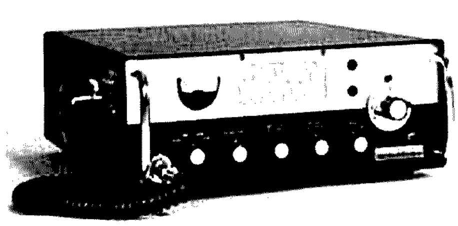
cause of the ability of modern-day electronic computers to handle a large quantity of information in a relatively short length of time. The computer selected the substitutes listed in this handbook in much the same manner that an individual would select a transistor replacement. The electrical and physical parameters shown in the manufacturer's published specifications for each bipolar transistor were given to the computer, and then each transistor was compared with all the others. Over one billion data comparisons were made in the preparation of this book.

The transistors which matched within given limits are listed as substitutions. A second section contains additional information on general purpose replacement transistors: the manufacturer, the polarity (npn or pnp), the material (germanium or silicon), and the recommended applications. The information in this handbook can be used by anyone concerned with transistor replacement – be it in amateur, industrial, commercial, or home-entertainment equipment. 160 pages, softbound. $2.25 from Comtec Book Division, Box 592, Amherst, New Hampshire 03031.

marine radiotelephone



For shipboard and shore applications RF Communications has produced a marine single-sideband radiotelephone incorporating the latest electronic advances and compatible with all existing and future marine stations. The new RF-201M series includes compatible a-m, speech clarifier, squelch, balanced audio and complementary accessories including remote controls, 1-kW amplifier, antenna couplers and power supplies. The unit is

offered in six basic models with up to 60 channels for ship and shore applications.

This two-in-one radio eliminates the need for a separate a-m radio by providing the equivalent talk power of 75 watts when operating in its compatible a-m mode. The rated ssb output power of the RF-201M is 150 watts PEP, more than enough to provide highly dependable long-range ssb communications. Transistor circuitry is used in all but the power amplifier section to provide ultra-high reliability with efficiency to keep power source drain low. Literature, further information and full specifications are available from National Sales, RF Communications, Inc., a subsidiary of Harris-Intertype Corporation, 1680 University Avenue, Rochester, New York 14610, or use *check-off* on page 126.

73 vertical, beam and triangle antennas

This new book by Edward M. Noll, W3FQJ, describes 73 individual vertical and beam antennas, beginning with simple construction and progressing to more complex arrangements. All you need are telescoping masts, wires, insulators, tubing, ingenuity and a desire to experiment. Each antenna described in this book was constructed by the author without assistance.

Antenna types range from simple dipoles, through verticals and yagis, to quad and triangle beams, and the topics are arranged in a sequential manner. The necessary mathematics are included but no extensive knowledge is required to build the antennas described. Simple test instruments are shown which will enable you to optimize the designs and obtain maximum antenna performance. Many of these antennas compete with, and some surpass, the performance of commercial beams.

160 pages, softbound. Published by Howard W. Sams & Co., Inc. $4.95 from Comtec Books, Box 592, Amherst, New Hampshire 03048.

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time-standard receiver

A compact new receiver, the model *STR-1* standard-time receiver, is now available from Caringella Electronics. The *STR-1* receives continuous standard-time and standard-frequency broadcasts from WWV on 5, 10, and 15 MHz. Optional coverage is also available for Canadian standard-time broadcasts from CHU on 7.335 and 14.670 MHz.

Ideal for a number of applications, the receiver can be used by industrial labs, radio and television stations, two-way radio service centers, radio amateurs, astronomers, boating and sports car enthusiasts, as well as others interested in accurate time or frequency.

Operation is simplified; only the volume control with on-off switch and the channel selector are found on the front panel. You simply turn on the receiver and select the frequency with the strongest signal; no need to hunt for the signal since each channel is crystal controlled. A choice of three frequencies assures 24-hour reception anywhere in the United States.

Sensitivity is 0.25 μV for 10 dB signal-plus-noise to noise. Dual-gate mosfets are used in the front end to achieve high sensitivity, low noise and good agc characteristics.

Ten transistors, two silicon diodes, one germanium diode and one zener diode are used in the circuit. The *STR-1* operates from and ac line or from an internal 12-volt battery.

The *STR-1* receiver is available in kit

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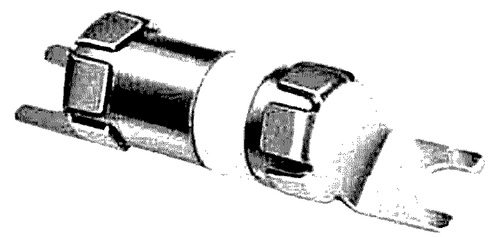
form, complete with all parts and step-by-step instructions. No test equipment is needed for alignment. Kit price is $74.95; factory wired and tested units are $94.95. A free data sheet, complete with technical specifications and schematic diagram, is available on request. For further information contact Caringella Electronics, Inc., Post Office Box 327, Upland, California 91786, or use *check-off* on page 126.

surge voltage protectors



New power-type *Fail-Safe* gas-filled surge voltage protector tubes have been developed by Siemens to protect circuits and equipment against lightning, power surges and other transients. Fail-safe is a term used to describe the protector's ability to provide a permanent short when it is subjected to extraneous current exceeding its discharge capability. By shorting, the SVP prevents the transient from destroying equipment. Additional parts, such as special holders or heat-sensitive components, are not required.

In addition to being inherently fail-safe, the Siemens S8-C350 tube has an ultrafast response time with a DC striking voltage of 200-400 volts, insulation resistance greater than 10^{10} ohm, 20,000 amps rated discharge capability and capacitance less than 3 pF. At extremely high surge currents of up to 50,000 amps, the SVPs still remain operative. Typical applications are protection of communication lines, power lines, power supplies and instruments. For more information use *check-off* on page 126 or write to Siemens Corporation, 186 Wood Avenue South, Iselin, New Jersey 08830 and request product bulletin SCD-1170-S100.

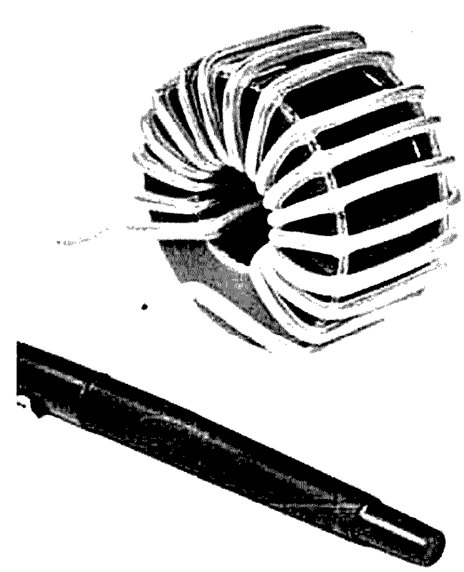
new motorola hep devices

A total of 167 new solid-state devices have been added to Motorola's HEP semiconductor line, expanding the selection to 470 devices. The 1971 HEP introductions cover the full range of devices including light-emitting diodes (LEDs), phototransistors, standard TTL and complex function TTL integrated circuits, high threshold logic (HTL) ICs, diode-transistor logic (DTL) ICs. In addition, high-speed emitter coupled logic (ECL) and linear devices have been added.

The diode end of the line has been expanded to include uhf hot-carrier diodes and voltage-variable capacitance tuning diodes, as well as 5- and 10-watt zener diodes. Other standard silicon transistors have also been added.

For more information use *check-off* on page 126 or write to Motorola Semiconductor Products Division, Post Office Box 20912, Phoenix, Arizona 85036.

toroid tank-circuit kit



The new toroid tank-circuit kit available from Redline Electronics contains all the parts required to build a highly efficient, compact pi-network inductor. Three basic circuit designs are provided with each kit; you can use the design that

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best fits your requirements for power and physical size.

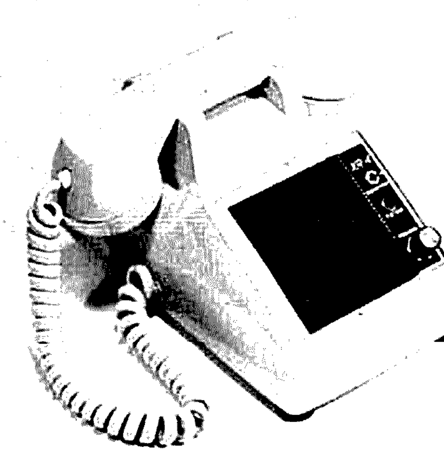
Design A is capable of handling 2 kW PEP; however, it should not be used if 1 kW continuous key-down operation is contemplated. Design B is electrically similar to Design A. However, it has an outboard 20-meter coil to reduce excessive core heating. This configuration will handle key-down inputs of 1 kW for prolonged periods.

Design C is capable of handling the greatest power (3 kW PEP) and is the most efficient design. An outboard coil is used for 10, 15 and 20 meters. The toroid is used on only 80 or 40.

All circuits are designed to be used with tubes which operate at voltages between 3,000 and 1,800 Vdc at peak current up to 1 ampere. For higher voltages and lower currents tap positions will require adjustment.

The no. 115 2-kW toroid kit contains a special 2-inch toroid core, end spacers, Teflon sleeve and wire plus complete instructions; $16.95 from Redline Electronics, 3498 East Fulton Street, Columbus, Ohio 43227. Also available are a bifilar filament choke with 30-amp rating, catalog no. 421, priced at $7.95; and 1.5-amp plate rf choke (2.5 to 55 MHz), catalog no. 417, priced at $6.95.

base-station control extension



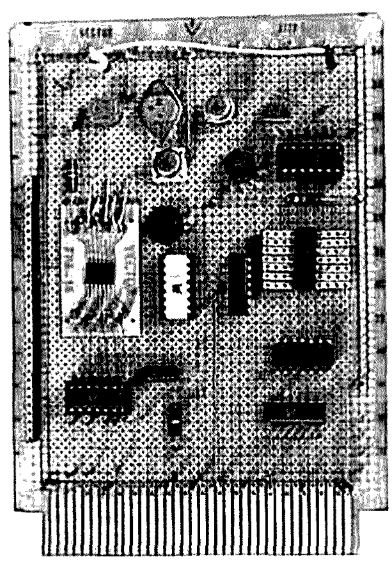
The new XR-4 remote extension from Alpha Electronics is designed to enhance the efficiency and convenience of your

repeater system. The unit features center button push-to-talk handset, a choice of colors, automatic monitor of the tone function, transmit indicator light and volume control. In addition, the XR-4 control extension is available with several options, including intercom system, multifrequency switching, handset volume control and choice of microphone types.

Of special interest is the provision that makes it possible to use the unit (with an add-on module) as a fully amplified remote on a telephone line. The XR-4 may be easily connected into any base station or dc remote through an 8-conductor cable. Cable length is limited only by line loss and receiver or remote amplifier output power; distances greater than 300 feet are often achievable.

The XR-4 base-station control extension is priced at $89.50 from Alpha Electronic Services, Inc., 8431 Monroe Avenue, Stanton, California 90680. For more information use *check-off* on page 126.

high-density plugboards



Vector Electronic Company has just announced a new high-density Plugboard series for mounting DIPs and discrete components. The Plugboards are made from glass epoxy or phenolic and are punched with an overall grid of .042-inch diameter holes located on 0.1-inch

More Details? CHECK–OFF Page 126

centers. The boards will accommodate any component with lead spacing on 0.1-inch multiples. As a result, integrated circuit packages or sockets may be mounted in any required density. Discrete components or hardware items such as test jacks and clips may be located anywhere on the board without fear of interfering with any pre-etched circuitry.

The new boards are offered with a choice of etched nickle and gold-plated contacts, Elco Varicons or Vector edge pins. Along with the Plugboards Vector is also supplying mating receptacles with a choice of solder-eye termination or square posts for wrapped wire connections.

The company also makes a full line of other accessories for the cards including rack-mounted cages and module cases for housing the boards, IC sockets, and various types of terminals which will fit the boards. Pricing of the new Plugboards ranges from $6.00 to $8.00, depending upon the type of board. For more information, use *check-off* on page 126, or write to the Vector Electronic Company, 12460 Gladstone Avenue, Sylmar, California 91342.

crystal cw filter

The latest addition to the crystal-filter line offered by Spectrum International is the new KVG XL10M, a ten-pole 500-Hz CW filter with superb skirt selectivity. Near-Gaussian response to -6 dB eliminates ringing; built-in matching transformers eliminate the need for external inductors.

The center frequency of the new crystal filter is 9.0 MHz; bandwidth at -6 dB is 500 Hz. Shape factor (6:60 dB) is 2. Insertion loss of the XL10M is 10 dB maximum; ultimate rejection is 80 dB minimum. The XL10M is priced at $59.95 from Spectrum International, Post Office Box 87, Topsfield, Massachusetts 01983. For more information on the XL10M as well as other crystal filters and crystal discriminators, use *check-off* on page 126.

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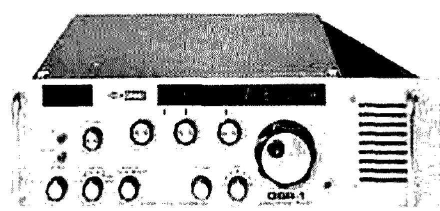
break-in cw system

The new *QSK MK II* is a station control system that permits you to hear other stations between your transmitted dots and dashes at speeds up to 60 wpm. The unit can be used with virtually any exciter/power-amplifier and receiver combination to provide high-speed antenna switching, exciter keying and receiver isolation. The *QSK MK II* will operate on all bands with no adjustment; the unit has been proven in traffic, contest and DX work.

The 8-transistor, 9-diode timing and sequencing circuit assures cold switching of the vacuum relay (5-kW rating). The unit features negligible losses and includes a built-in 117-Vac power supply. $119.95 from your dealer, or write to Dynamic Technology International, Inc., 8 Fellowship Road, Cherry Hill, New Jersey 08034. For more information use *check-off* on page 126.

communications receiver



The new Drake model DSR-1 receiver is a commercial grade communications receiver employing the most up-to-date solid-state devices and circuitry offering continuous coverage from 10 kHz to 30 MHz. The received frequency is indicated by six Nixie tubes to 100 Hz. All frequency injections are controlled by a phase-locked digital synthesizer which allows incremental frequency selection in 10, 1 and 0.1 MHz steps. The remaining zero to 0.1 MHz is continuously adjustable from the front panel by a highly stable variable oscillator. Modular construction on easily accessible printed-circuit boards is used throughout.

The large use of dual-gate mosfet

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transistors in the DSR-1 contributes to the receiver's superior intermodulation, avc, wide dynamic range and overload performance. The front panel controls allow the operator to select frequency, a-m or ssb product detector, i-f bandwidth, af gain, bfo pitch, fast or slow avc, manual rf gain and the highly effective Drake series-gate noise blanker. lsb (independent sideband) is a built-in feature of the DSR-1. Separate i-f crystal filter, i-f amplifier and audio output circuits allow two simultaneous communication channels to be used on one frequency assignment, doubling the information receiving capacity.

Consideration in the design of the Drake DSR-1 has been given to special customer requirements. Accessories and space have been reserved in the DSR-1 so that performance and operating modes to fill special customer needs can be accommodated. For more information use *check-off* on page 126, or write to the R. L. Drake Company, 540 Richard Street, Miamisburg, Ohio 45342.

high-voltage solid-state rectifiers

You can now replace a large selection of high-voltage mercury and vacuum tube rectifiers with Semtech's solid-state silicon-rectifier stacks. Designed to last the

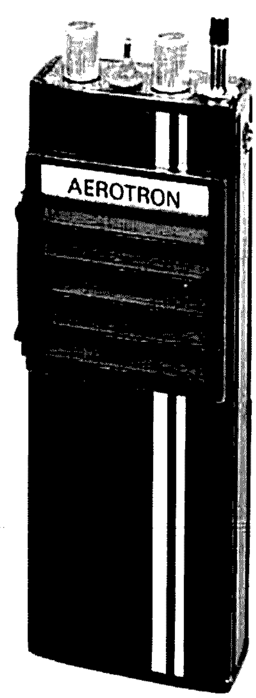
More Details? CHECK—OFF Page 126

life of the equipment, Semtech's stacks are smaller, corona free, dissipate heat more effectively and eliminate filament power. No special adaptors are required.

Most of the mercury and vacuum tubes used today are replaced every few months and are a constant source of irritation, down time and expense to the user. Semtech Stacks will replace such mercury rectifier tubes as 575A, 872A (pictured), 7014, 7018, 7019, 7019L, 673, 816, 866A, 5558, 5561 and 8008. Vacuum rectifier replacements are 8020, 5973, 2-450A, 576A, 6303 and 561.

Transmitters, induction heating units, welding equipment, electrostatic paint units, and amateur radio gear use high-voltage rectifiers. For more information write to Semtech Corporation, Newbury Park, California 91320, or use *check-off* on page 126.

portable vhf-fm radio

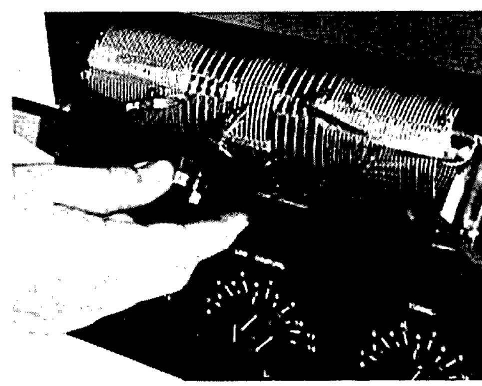


Aerotron, Inc. has announced the first of a series of shirt-pocket size, hand-held, two-way vhf-fm radio equipment. The new Aerotron "600" series is compact, durable and easily serviced. The unit features load and clear 1-watt audio for noisy environments, and is available in the frequency range of 146 to 174 MHz. Test jacks make possible all important

measurements such as rf power output, deviation, frequency, receiver sensitivity, audio output and distortion without the need to open the case.

The "600" series features high frequency crystal-lattice filtering pioneered by Aerotron in the land-mobile services, and is housed in a sturdy case. The transmitter section is available in power output levels of 100 milliwatts, 1.8 watts and 5.0 watts. Options include continuous-tone squelch, multifrequency capability (up to six channels) and industry-compatible continuous-tone squelch. For further details use *check-off* on page 126 or write to Aerotron, Inc., Post Office Box 6527, Raleigh, North Carolina 27608.

universal transmatch



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More Details? CHECK—OFF Page 126

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focus on communications technology . . .

ham radio magazine

NOVEMBER 1971

automatic radiotelegraph translator and transcriber

this month

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contents

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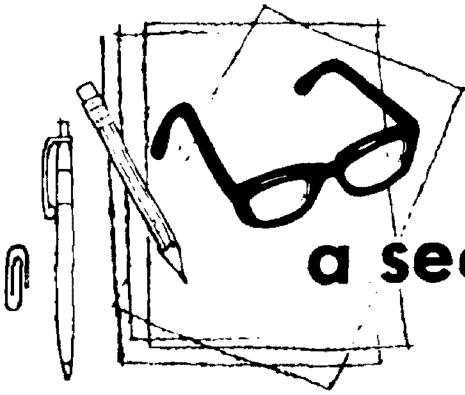


a second look by jim fisk

The editor of a technically oriented magazine like *ham radio* wears several hats. I could use this whole page to describe the details that require attention to keep the magazine running smoothly, but I'd like to talk for a moment about one very important editorial task that means the difference between an interesting, accurate technical magazine, and one that isn't.

The articles in *ham radio* are written by authors not unlike you, the reader. They range from enthusiastic hams who want to share an idea, to fellows with engineering backgrounds (who also want to share an idea). I welcome the output of anyone who is interested in contributing something which will benefit all hams.

Budding authors often ask, "What kind of articles are you looking for?"

That question is difficult to answer since many new manuscripts arrive every day, but generally speaking, I am looking for simple construction projects that the average reader can put together in one or two weekends. Larger construction projects are also welcome, but the average *ham radio* reader must split his leisure time between amateur radio and other interests, so he doesn't have time to build a Chinese copy of a complex piece of gear.

When I read an article contributed to *ham radio*, the first thing I look for is interest value. If the manuscript passes this test, the next thing I look for is technical accuracy and attention to detail.

The contributed article doesn't have to be a literary masterpiece. If you have a good idea; if it's well documented; if the illustrations and technical discussion are clear and accurate – you may have a winner!

Don't feel too badly if your article is not accepted. Since we publish about twelve feature articles in every issue, to keep the production pipeline full, we purchase that same number of new manuscripts each month. This keeps the article backlog to a minimum and insures that the fare served up in each issue of *ham radio* represents the latest possible information on any given subject.

During an average month I receive about 45 or 50 new manuscripts. At some point during the month I sit down and go through each of the articles, rejecting those that are clearly not usable in *ham radio*; this narrows the stack of manuscripts down to 25 or so.

Now comes the most difficult part – deciding which of those remaining 25 articles are most desirable. Articles that are too long or need more polishing are returned to the author with suggestions for rewriting the article to our standards. The remaining material is further screened for technical accuracy and reader interest; this process continues, with comments from the staff (and sometimes, arguments), until we have decided upon the articles which will be published in *ham radio*.

If you prefer to read *ham radio*, rather than write for it, you can help by telling me the kind of article you like. If you have a pet project in mind, or an old project that could be updated with transistors or ICs, let me know about it; I'll pass the idea along to one of our regular contributors.

We are continually trying to improve *ham radio*, but we can only do that successfully if we hear from you, the reader, so keep those cards and letters coming in.

Jim Fisk, W1DTY
editor

publisher's log

skip tenney
W1NLB

There seems to be a never ending discussion among both amateurs and those who cater to the amateur fraternity with goods and services as to whether or not amateur radio is growing. Most of the conclusions recently have been rather negative.

However, an interesting new twist has recently been introduced into arguments about this growth, or lack of it.

If you look at the total operator license figures from the FCC, they show a virtually stable figure since 1965, although growth was relatively constant until that time.

Does this mean that our ranks are not growing any longer? A number of people will disagree with this idea. Remember that, in 1964 a $4.00 fee was introduced for all but Novice licenses, while in 1969 this fee was increased to $9.00.

Unquestionably, there are former licensees who had been continually renewing their ticket when it came up for renewal every 5 years even though they never used it. If it was free – why not? The four-dollar fee probably dropped some of these people while the nine-dollar license stopped a good many more.

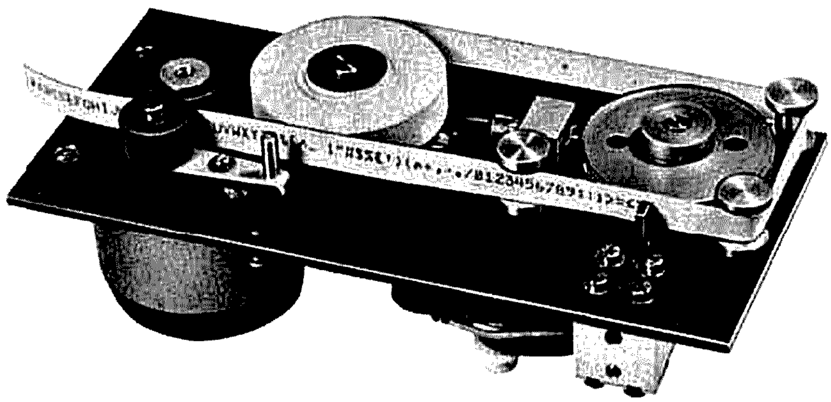
As these former licensees drop out, we are not losing active amateurs. They were amateurs in name only, but the new licensees who have come along to replace them are far more apt to put that shiny new ticket to work on the air.

Thus, the argument can logically be made that although the total of licenses outstanding has not grown, the actual number of active on-the-air amateurs has continued to increase.

This is a hard theory to prove or disprove because there have been many changes in our operating habits in the last ten years. Sideband has virtually replaced a-m on most high-frequency phone allocations, thus, permitting greater use of these frequencies. The explosive growth of vhf fm has also absorbed much activity in previously seldom-used portions of the amateur spectrum. Therefore, you cannot easily evaluate activity just by listening on our bands. We are using them more efficiently and are making room for added activity without necessarily having greater crowding (except possibly in a good pile-up on 20 meters).

The answer to the problem of growth is not found as simply as this editorial might imply, but it does show that amateur radio may be doing a good deal better than some people think.

Skip Tenney, W1NLB
publisher



automatic radiotelegraph translator and transcriber

This automatic fist follower converts the CW output from your receiver into the printed word

Clarence Gonzales, W7CUU, Richard Vogler, K7KFA

In the past, amateur radio operators have devoted a lot of effort to making it easier to send perfect (or near perfect) CW signals. These efforts have progressed from the straight manual hand key to the semiautomatic bug, the electronic keyers, and finally, to the keyboard sender.[1] However, reception has been left to the ability of the human operator.

There have been a few techniques devised to aid the receiving operator, but most of these have been used to copy high-speed machine-sent Morse. One technique produces an inked recording of the high-speed CW transmission; the recording is fed through a photocell arrangement at greatly reduced speed to produce an audible tone that is translated by a human operator. Magnetic tape may be used in much the same way, but a human operator must still be used to translate the radiotelegraphic signals into the written word.

The idea of automatically translating and transcribing Morse code without a human operator is not new. Some time ago researchers at the Massachusetts Institute of Technology used a digital computer to recognize hand-sent Morse characters.[2] The computer program designed for this task was called MAUDE for Morse Automatic Decoder.

In the MAUDE program each *mark* and *space* was assigned a number that represented its duration. Then a sliding-window algorithm was used by the computer to identify the types of *spaces* and types of *mark* signals. The program provided for continuous adjustment based on the last six elements of each type; this resulted in automatic tracking of variations in hand-sent radiotelegraph signals. With this approach misinterpretation occured only about once in 10,000 times.

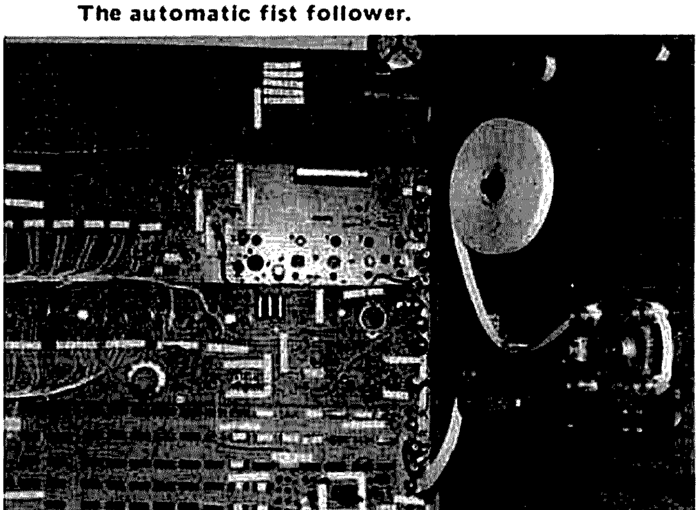
Although the MIT studies showed the feasibility of automatically translating hand-sent Morse code, unfortunately, the average radio amateur does not have his own digital computer for such activity.

automatic fist follower

The most practical and economic approach to automatic translation of Morse code would be to design a special purpose computer just for this function. The automatic fist follower (AFF) described in this article is just such a device. Although it uses an entirely different algorithm for its operation, it performs nearly as well as the MIT MAUDE computer system.

The automatic fist follower was originally conceived after reading Horowitz's article on the keyboard CW sender. However, we had to wait for the advent of inexpensive integrated circuits to make it technically practical and economically feasible for the amateur.

The automatic fist follower.



Nevertheless, the design was quite a challenge, and previous experience with computer logic design didn't hurt a bit. The ICs used in the AFF were scrounged from castoff vendor's samples and surplus computer circuit-boards. The discrete parts — transistors, diodes, resistors, capacitors — came from the junk box. The one item which had to be purchased new at considerable expense was the strip printer.

The automatic fist follower takes the audio output of a communications receiver that is tuned to a radiotelegraphic signal and translates it into letters, numerals, special symbols and spaces which are transcribed onto a paper tape by a strip printer.

The AFF will accept any code speed between 5 and 60 wpm. It would be a simple matter to extend the range to 300 wpm (the maximum speed of the strip printer) by increasing the frequency of the AFF's internal clock. However, in order to use a fairly large time constant in the signal filter to obtain good noise rejection the limit was set at 60 wpm.

The automatic fist follower will automatically track any reasonable variations in sending speed, dot length, dash length, weight, etc. In many cases it will track sending variations that are beyond the reception capability of a good human operator. All the operator has to do is tune in the sending station, set the approximate wpm control on the AFF and sit back and read the copy coming on the printer.

basic detection theory

Before discussing the functional block diagram of the automatic fist follower, it is desirable to explain the theory behind the detection of widely varying dots, dashes and spaces with which the AFF must cope. Of course, the easiest task for the AFF would be the reception of machine-sent Morse code, since this would be nearly perfect and free from variation with strict adherence to the standard definitions of dot length, dash length, element space, character space and word space. For example, if a dot is

taken as a unit length, the standard definitions prescribe a length of 3 units for a dash, 1 unit for an element space, 3 units for a character space and 7 units for a word space.

Human operators *roughly* approximate perfect code; some operators deviate so far no one can interpret most of what they send. A convenient way of describing variations from perfect code is by a parameter called weight. Weight is defined as *dot length* divided by *dot length plus element-space length.* Perfect code has a weight of 50%:

$$\text{weight} = \frac{\text{dot}}{\text{dot} + \text{space}} = \frac{1}{1+1} = 50\%$$

Light code has weight less than 50%; *heavy* code has weight greater than 50%.

The AFF uses a sampling technique to determine whether it is receiving a dot, a dash or a space element. Both the *mark* (dot or dash) and *space* (interval between *marks*) are sampled with a series of clock pulses. The number of clock pulses which occur during the *mark* signal are counted. Likewise, the number of clock pulses which occur in the interval between *mark* signals are counted. The frequency (wpm) control on the AFF is initially set so that most of the dots give a *mark* count of 6. The operator is assisted in this adjustment by the dot indicator light which flashes quite consistently when the setting is correct.

The typical dot will give a count of 6, the typical dash a count of 18, the element space a count of 6, the character space a count of 18 and the word space a count of 42 (more if the last word was transmitted). These counts may vary by a large margin and still be recognized by the AFF. Any *mark* count between 1 and 12 is recognized as a dot. A *mark* count of 13 or more is treated as a dash. A *space* count between 1 and 12 is an element space, a count between 13 and 30 is a character space and any count of 31 or more is treated as a word space.

It is interesting to calculate the weight value limits which can be received and properly interpreted by the AFF. The lightest code it can handle is a dot length

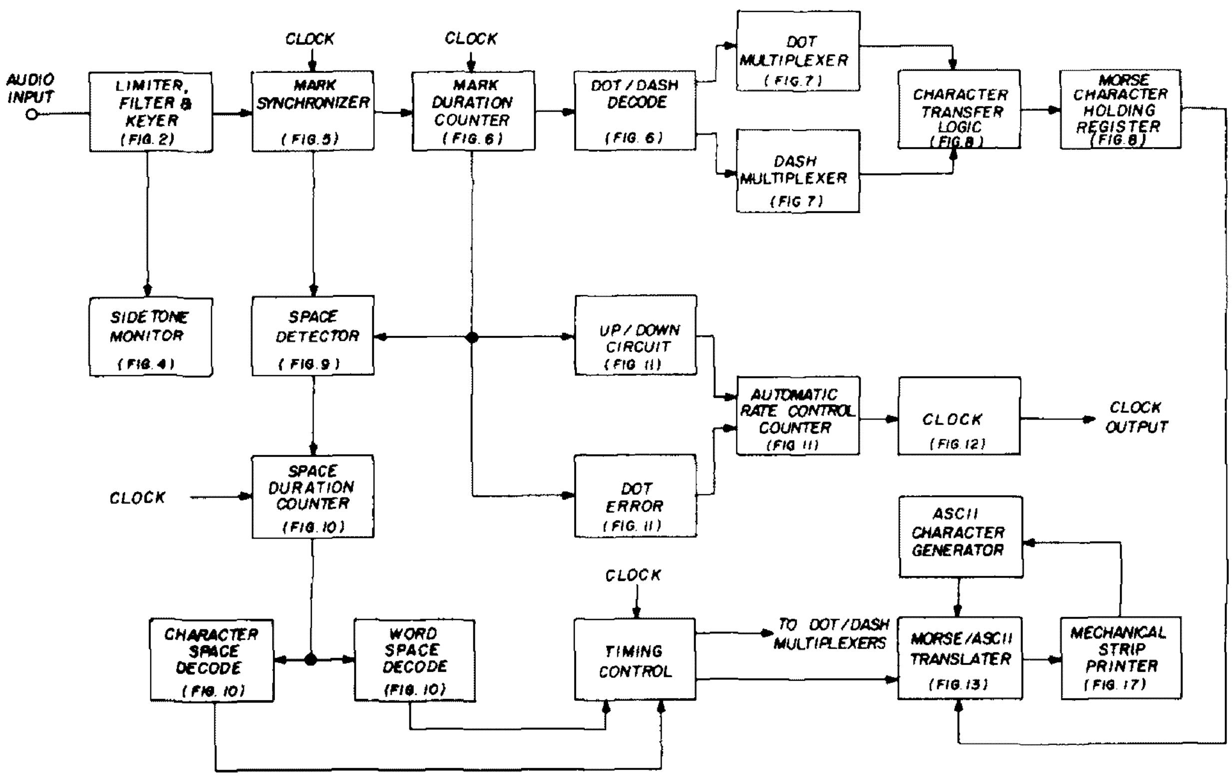


fig. 1. Block diagram of the automatic Morse code translator and transcriber.

of 1 count and an element space length of 12 counts, resulting in:

$$\text{weight} = \frac{1}{1 + 12} = \frac{1}{13} = 7.7\%$$

The heaviest code the AFF can handle is a dot length of 12 and an element space length of 1, resulting in:

$$\text{weight} = \frac{12}{12 + 1} = \frac{12}{13} = 92.3\%$$

However, the AFF's automatic-frequency-control circuits will allow even greater variations in sending characteristics than this to be recognized. If the dot length varies from a 6-clock pulse count, a dot length error condition is detected and the frequency is automatically adjusted in a direction to bring the clock pulse count back to 6. Therefore, the AFF follows gradual changes in sending speed and will actually track the sender's fist through a wide range of variations.

theory of operation

The audio output from a communications receiver is fed into a limiter/clipper (see fig. 1). The radiotelegraph signal is shaped, and in successive circuits it is filtered and amplified so that it is suitable to trigger the keying circuit. The output of the keying circuit, which follows the on-off characteristics of the original signal, is synchronized with the AFF clock by the mark synchronizer.

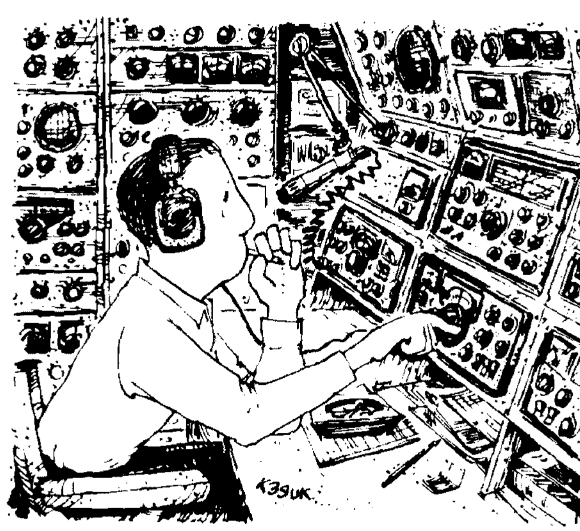
The synchronized *mark* signal is used to control the start and duration of a count in the *mark* duration counter. This counter effectively counts clock pulses throughout the duration of the *mark* signal. At the end of the *mark* signal the count is terminated and a decision is made by the *dot dash decode* circuit as to whether the element is a dot or a dash. The appropriate indication is stored in the first stage of either the dot or dash multiplexer, and the *mark* duration counter is reset.

Each multiplexer is a five-stage shift register which performs the basic function of serial-to-parallel conversion. In other words, each element (dot or dash) of the character is successively stored in the multiplexer until all of the character has been received. All of the character elements are then available for translation prior to printout.

In addition, at the end of the *mark* signal, the space detector turns on the space duration counter which counts clock pulses throughout the duration of the *space* (i.e., until either the next *mark* signal begins, or until the counter has reached its maximum count, signifying a word space). At the completion of the *space* interval, the count is terminated and a decision is made by the character and word space decode circuits as to whether an element space, character space or word space has occurred.

If an element space is detected the mark duration counter again counts clock pulses during the next *mark* signal, and the space duration counter is reset. If a character space is detected the timing control circuit causes the dot/dash data in the multiplexer to be transferred to the Morse character holding register through the character transfer logic. The multiplexers are then reset for the next character.

If a word space is subsequently detected the timing control causes a *space* to be transferred to the Morse character holding register. In either case the next



SSO . . . SSO . . . SSO . . .
I wonder what that means?

mark signal causes the space duration counter to be reset.

As long as only element spaces are detected the dot/dash data continues to be shifted into successive stages of the dot/dash multiplexers. It takes the detection of a character space to cause the transfer of this accumulated data into the Morse character holding register.

Once the character is in the Morse character holding register the timing control supplies a character-ready signal which activates the Morse/ASCII code-translator. This translates the Morse representation of the character into the corresponding ASCII* representation which is used by the strip printer. The actual character passing under the print hammer is followed by the ASCII character code generator which is a counter that is sychronized with the positions of the characters on the print wheel. When the character under the print hammer corresponds to the translated Morse character the timing control activates the print hammer, thereby impressing the character on a paper strip located between the hammer and the wheel. The paper then moves to the next position for the next character. The Morse character holding register retains the old character until a new character is transferred into it.

*ASCII is the American Standard Code for Information Interchange. This is an eight-level code that is commonly used in computer printout devices and some teleprinters.

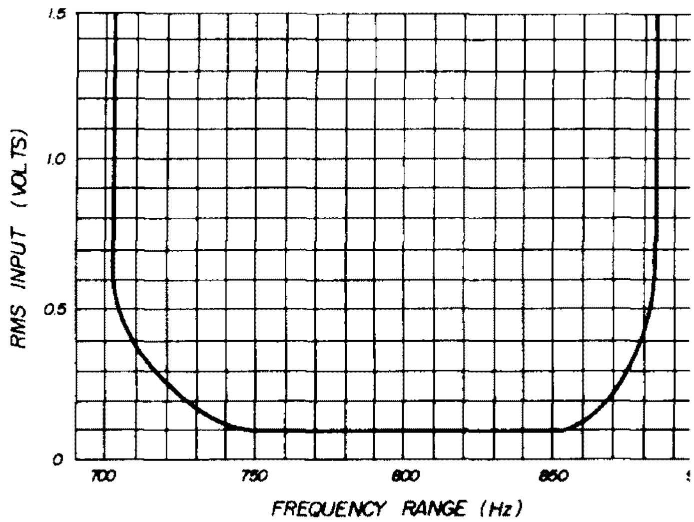


fig. 3. Bandpass characteristic of the 800-Hz active filter in fig. 2.

automatic frequency control

During the detection of dot information the dot error circuit indicates whenever a recognized dot element deviates

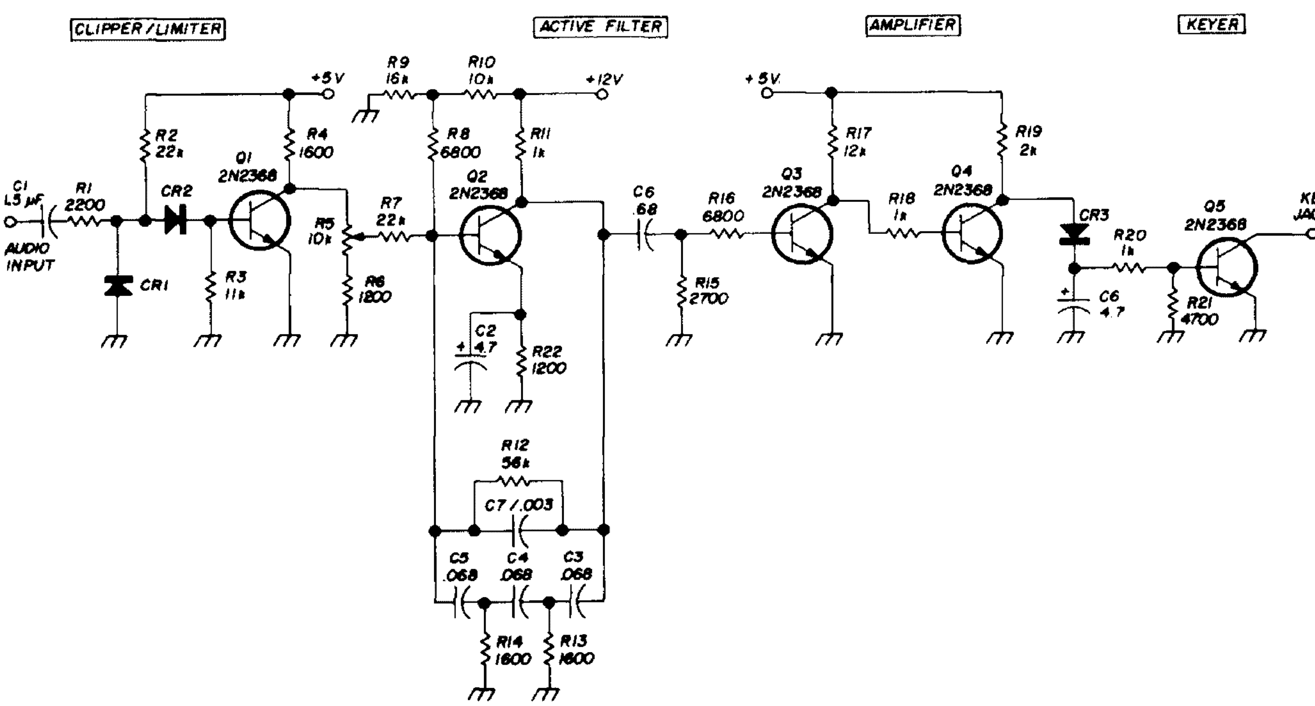


fig. 2. Filter and keying circuit. R12 and C7 provide negative feedback for improved filter skirt response.

from a perfect dot length of 6 counts. This error indication, in conjunction with the up/down circuit, causes the automatic rate control counter to either increase or decrease its count depending on the magnitude and direction of the dot error. The output of this counter is transformed into an analog signal by the analog error generator. This signal alters the AFF clock rate through the frequency control circuit so a dot generates a count of 6 in the *mark* duration counter. With this automatic following feature it is possible to track gradually changing sending speeds, up and down, to the limits of the clock frequency variation.

keying circuits

To translate radiotelegraphic signals with an automatic system the electronic circuits must be designed to perform the same elementary functions as a human operator to extract intelligence from the dot-and-dash impulses. It is a unique ability of the human ear and mind to concentrate on a desired CW signal and ignore static and interference.[3]

Although the circuits used in the AFF do not give complete immunity to static they do a good job of ignoring intruding voice and CW signals. With static down as little as 6 dB from the desired CW signal the AFF circuits efficiently imitate the performance of a human operator.

Although many single-stage circuits have been designed for audio frequency filters their overall performance is not optimum for the AFF. The circuit in fig. 2 uses five stages to provide the needed functions of selecting the desired signal over a wide range of input signals (0.2 volts to 100 volts rms) along with the required frequency selectivity of less than 10 Hz to 1 kHz. Since complete theory of operation can be found in any textbook on selective audio filters only the general operation will be considered here.

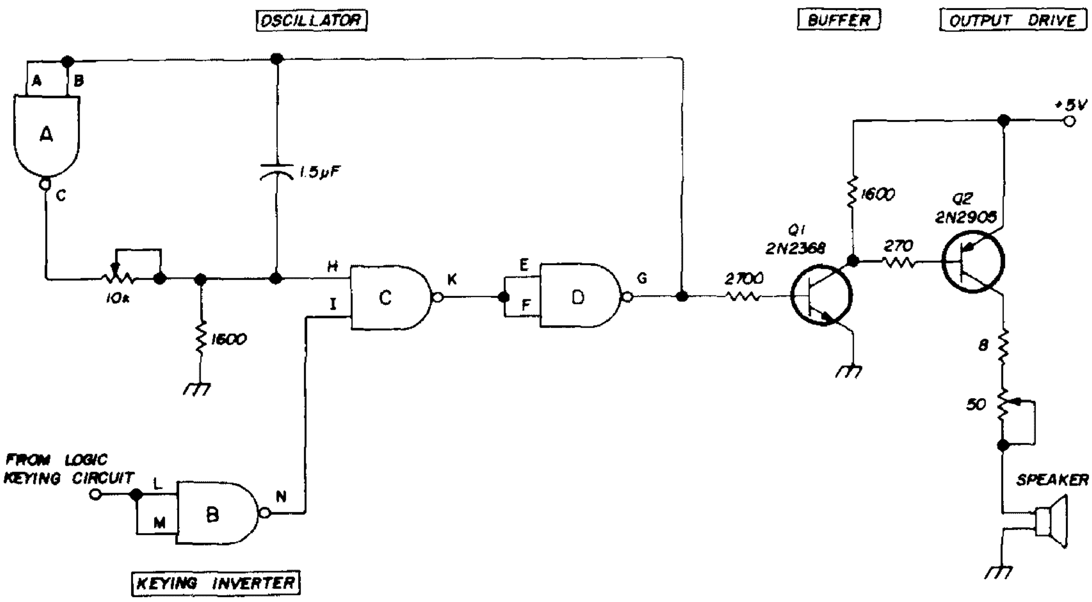


fig. 4. Sidetone generator. Suitable integrated circuits are Motorola SC2631, TI 16223 or Sylvania SG-220-07.

circuit operation

Capacitor C1 blocks dc voltage from the input transistor Q1 yet allows the audio frequencies normal to amateur receivers to pass through. Diodes CR1 and CR2, along with resistor R1 clip and limit the input signals to Q1 between about -0.7 to +1.4 volts; R2 and Q1 biased on. All input signals cause the collector of Q1 to swing from near zero to +5 volts to provide constant-level signals to drive the 800-Hz center-frequency active filter.

Resistors R5 and R6 adjust the signal level (and bandwidth) to the base circuit of Q2. All signals at the base of Q2 are inverted 180° by Q2. Phase-shift elements

C3, C4, C5, R13 and R14 cause an additional shift of 180°; this signal is fed in phase with the input signal to the base of Q2, causing the circuit to oscillate. Signals not at the filter's operating frequency are out of phase and will not cause oscillation.

Transistors Q3 and Q4 amplify the low level output signal from the filter, shaping and squaring it to drive the keying detector diode CR3. The detector circuit has a charge time constant of about five milliseconds, requiring around five cycles of operation from the active filter before the base of transistor Q5 can be biased on. This provides some immunity to static.

side-tone monitor

The side-tone generator (**fig. 4**) consists of a quadruple two-input positive TTL NAND gate driving two transistors. Three sections of the IC are connected as a three-stage ring oscillator; the fourth gate is used to invert the keying circuit which is grounded when keyed. Timing is controlled by the 10,000-ohm pot and the 1.5-μF capacitor. The squarewave generated by this circuit drives the amplifier transistors in the speaker circuit.

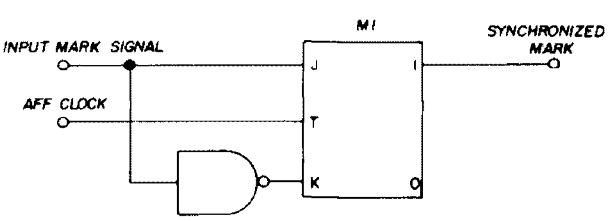


fig. 5. Mark synchronizer circuit.

mark synchronizer

The mark synchronizer (**fig. 5**) is merely a JK flip-flop which synchronizes the input signal from the keyer with the internal clock of the AFF. Synchronization is necessary to assure proper subsequent timing relationship within the AFF.

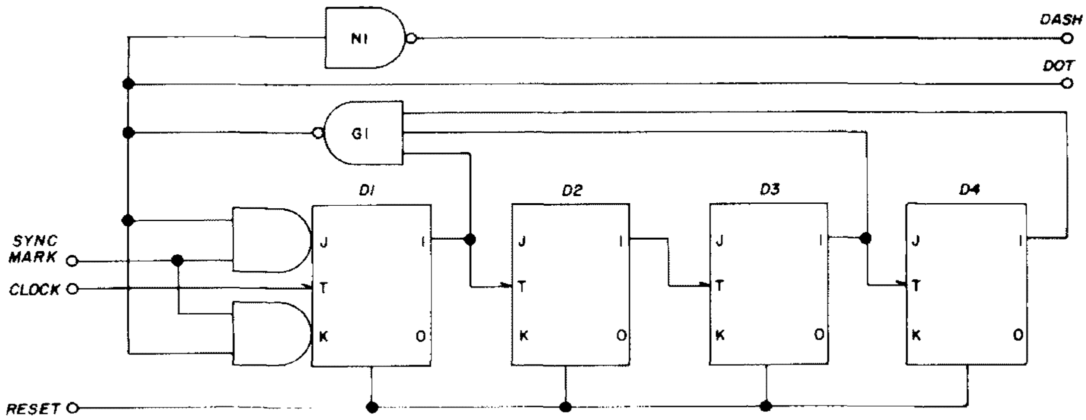


fig. 6. Mark duration counter and dot/dash decoder.

When the input *mark* signal goes positive the trailing edge of the next clock pulse sets the flip-flop to the *one* state. The flip-flop remains in this state until the input *mark* signal goes back to zero. At this time the trailing edge of the following clock pulse resets the flip-flop to *zero*. Thus, the flip-flop follows the input *mark* signal in synchronization with the AFF clock.

mark duration counter

Four JK flip-flops are used to form a sequential binary counter (**fig. 6**). This counter is used to count the number of AFF clock pulses which occur during the presence of a synchronized *mark* signal; the total count is indicative of mark signal duration.

When the synchronized *mark* signal goes positive the trailing edge of each successive clock pulse causes stages D1 through D4 to count up. When, and if, the counter reaches a count of 13 (D4, D3, D2, D1=1101), the output of gate G1 goes to zero and freezes the counter at

this total. The counter is reset when the *mark* signal goes back to zero and after a decision is made as to whether the total count represents a dot or a dash.

You may recall that any mark count up to and including 12 is detected as a dot. Any count of 13 or more (always represented by a counter state of 13) is detected as a dash. The adjustable oscillator is initially set so a typical dot will give a count of six.

Note that when the binary counter has terminated its count the output of gate G1 (when positive) indicates a dot. The output of inverter N1 (when positive) indicates a dash.

dot and dash multiplexers

When a dot/dash decision has been made by gate G1 and inverter N1 at the conclusion of a *mark* count some means must be provided to remember this decision. This is accomplished in the case of a dot by the dot multiplexer; in the case of a dash the dash multiplexer is used. Each of these multiplexers consists of a 5-stage shift register composed of JK flip-flops (see **fig. 7**).

After the decision is made as to whether the previous *mark* signal is a dot or a dash a shift pulse is generated and stores this indication by storing a one in flip-flop T1 if a dot and in flip-flop H1 if a dash. If the next *mark* signal is part of the same character the following shift pulse stores this indication in T1 or H1 as appropriate. At the same time it shifts the previous indications in T1 and H1 into T2 and H2, respectively. Likewise, the previous indications in T2, H2 are shifted into T3, H3; those into T4, H4, and into T5, H5.

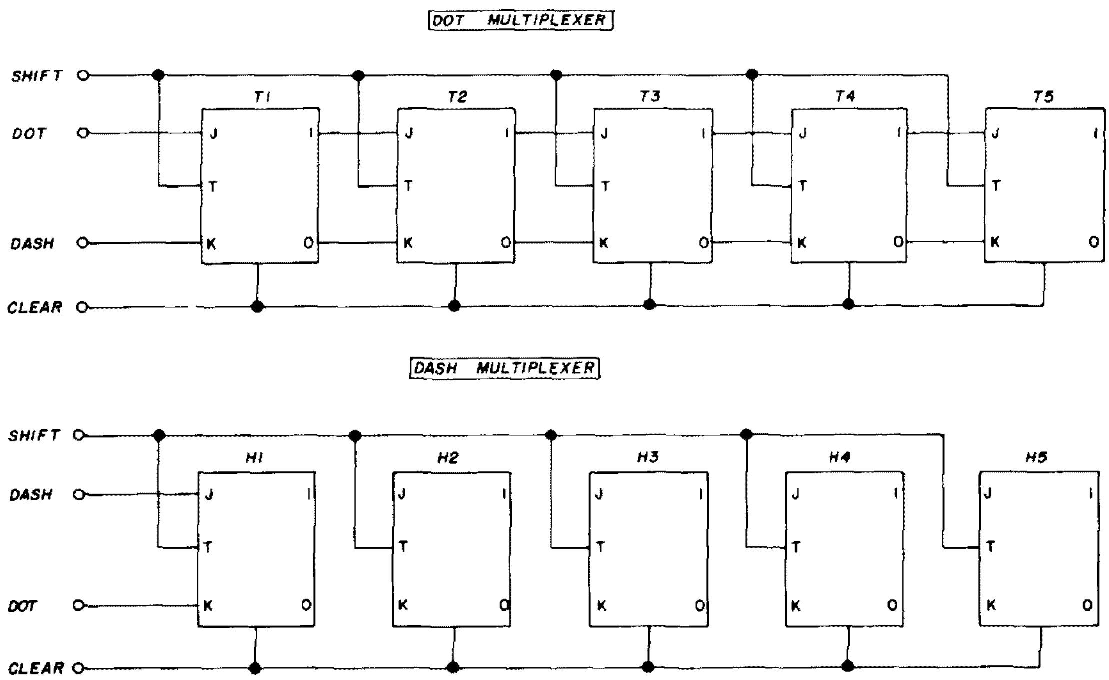


fig. 7. Dot and dash multiplexer circuits.

Up to five successive dot/dash character elements can be stored in the multiplexers. If a character is composed of six elements (e.g., the special symbols), the first element is shifted out of T5, H5 and lost at the time that the sixth and last element is shifted into T1, H1. However, the special symbols can still be decoded since a chopped-off five element representation of them can be uniquely identified.

A flip-flop is capable of representing two states and could therefore represent either a dot or a dash. However, there are actually three states to be represented: dot, dash and null (neither). As an example, suppose that a Morse code "W" is

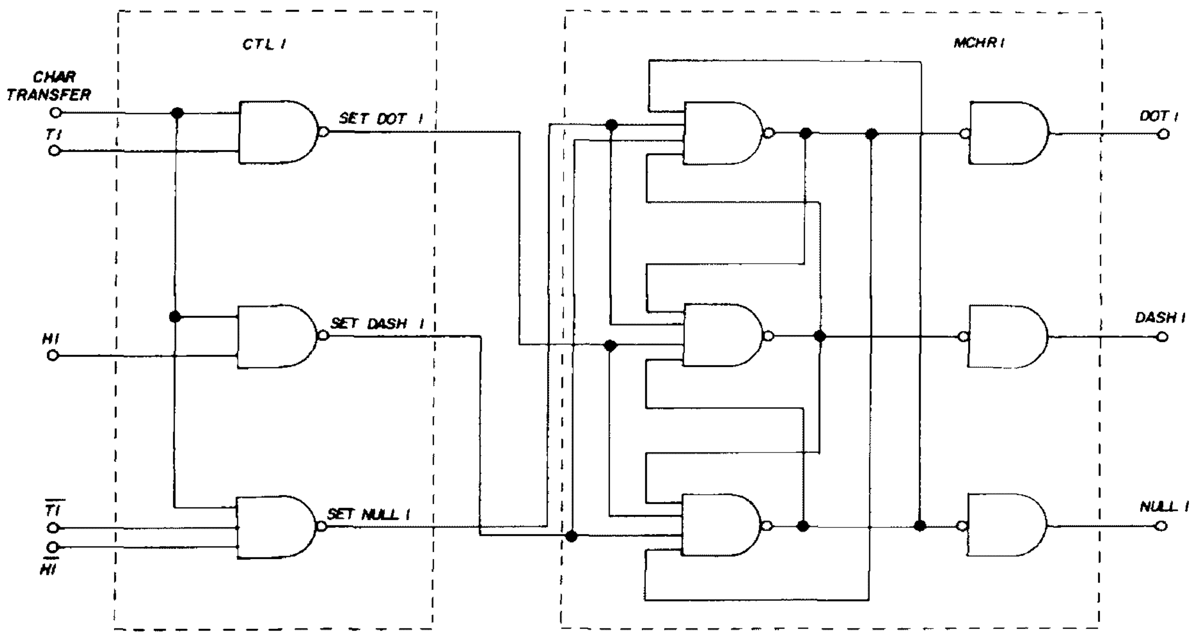


fig. 8. Character transfer logic (CTL) and Morse character holding register (MCHR).

stored in the multiplexers. In the dot multiplexer T1 is zero (reset), T2 is zero, T3 is one (set), T4 is zero and T5 is zero. In the dash multiplexer H1 is one, H2 is one, H3 is zero, H4 is zero and H5 is zero. This is interpreted as follows:

null is stored in stage 5 since both T5 and H5 are zero (reset)

null is stored in stage 4 since T4 and H4 are zero

dot is stored in stage 3 since T3 is one (set) and H3 is zero

dash is stored in stage 2 since T2 is zero and H2 is one

dash is stored in stage 1 since T1 is zero and H1 is one

(Stage 1 always contains the last element of the Morse character.) Note that it is not permitted to set both flip-flops in the same stage (such as T1=1 and H1=1) since that would mean that the same element was indentified as *both* a dot and a dash. The dot/dash multiplexers are cleared after their data is transferred to the Morse character holding register upon the occurrence of a character space.

morse character holding register

There are five stages comprising the character transfer logic and Morse character holding register (fig. **8**). This illustration shows a typical stage (in this case, stage 1) made up of a portion of the character-transfer logic (CTL1) and a three-stage latch (MCHR1) which holds element number 1 from the dot-dash multiplexers.

When a character space is detected the transfer-character pulse is generated which transfers the dot-dash data in the multiplexers to the Morse character holding register. The multiplexers are then cleared so they are ready to receive the next character. The MCHR holds the character until it can be translated and printed out.

It can be seen from fig. **8** that if T1 (dot indication) is present at the time of the transfer character pulse the CTL1 circuit causes the *set dot-1* output to go negative, causing MCHR1 to latch in the *dot-1* state. That is, the *dot-1* output goes positive while the *dash-1* and *null-1* outputs go negative. Similarly, if H1 (dash indication) is present during the transfer pulse, MCHR1 latches in the *dash-1* state. If neither T1 nor H1 is present during the transfer pulse, MCHR1 latches in the *null-1* state.

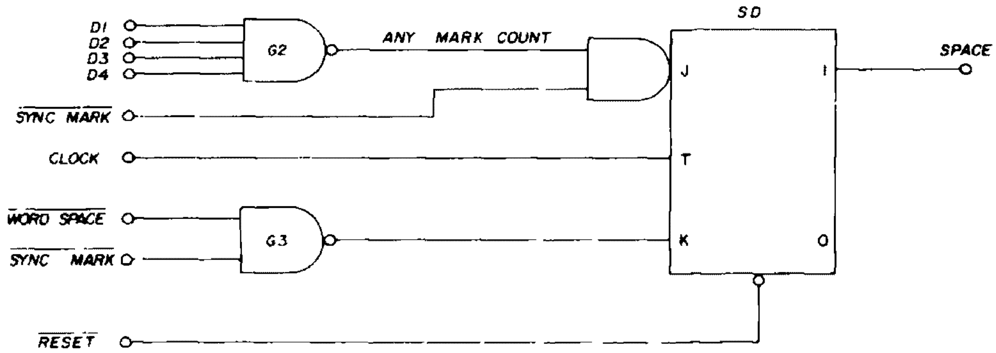


fig. 9. Space detector.

The outputs of MCHR1, together with the outputs of the other four stages of the Morse character holding register, are connected to the input of the diode matrix which performs the Morse to ASCII character translation. This must be done to get the character in a form suitable for printout.

space detector

The space detector (fig. 9) is a flip-flop that is set at the beginning of each interval between *marks* and reset at the beginning of the next *mark* if this *mark* occurs before a word-space is detected. If not, the space detector is reset upon the detection of the word-space.

Flip-flop SD is initially in a reset state. If the *mark* duration counter has any count in it the output of gate G2 is positive. At the completion of the *mark* signal *sync mark* goes positive enabling the set side of flip-flop SD. On the next clock pulse SD is set. The flip-flop will remain set (signifying a *space* condition) until the output of gate G3 goes high. This is caused either by the occurrence of a new *sync mark* signal or the detection of a word-space. If either of these conditions should occur flip-flop SD is reset on the next clock pulse. This signifies the end of the space interval.

space duration counter

The space duration counter (**fig. 10**) is a 5-stage serial binary counter composed

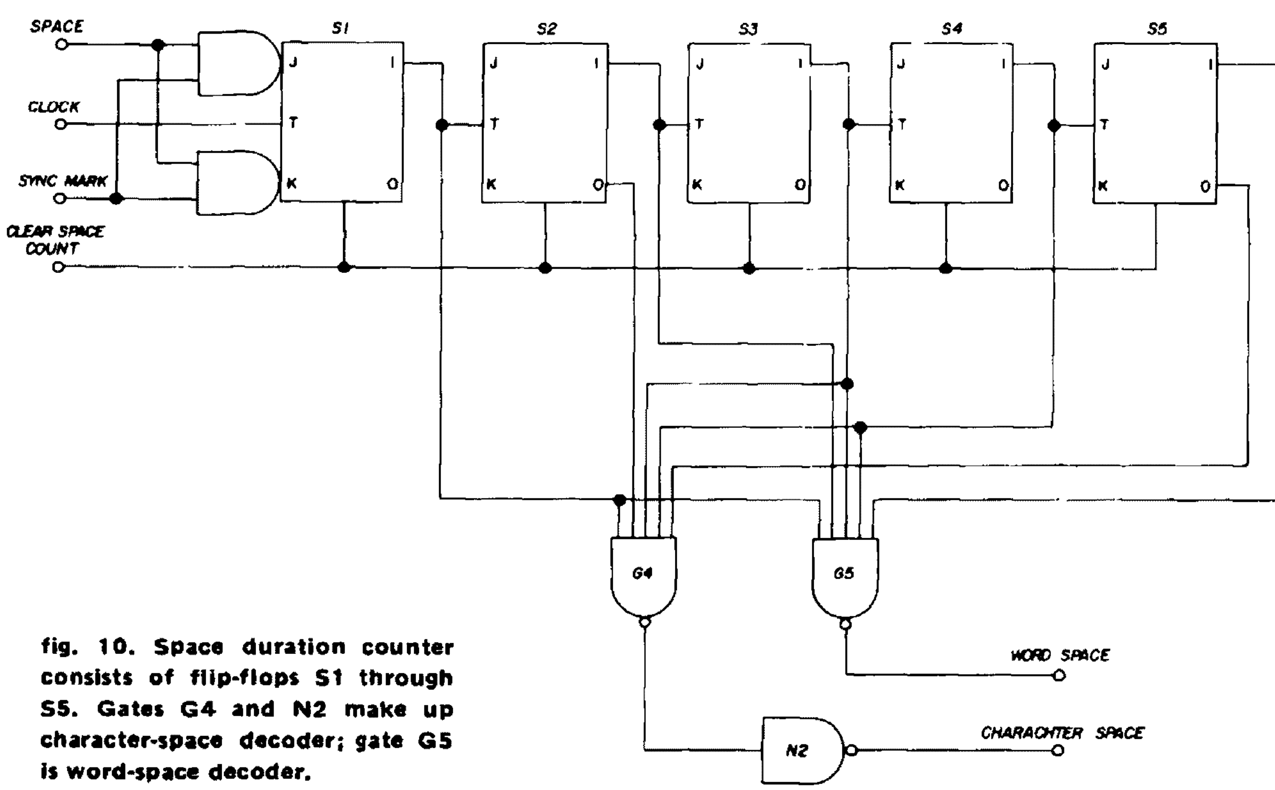


fig. 10. Space duration counter consists of flip-flops S1 through S5. Gates G4 and N2 make up character-space decoder; gate G5 is word-space decoder.

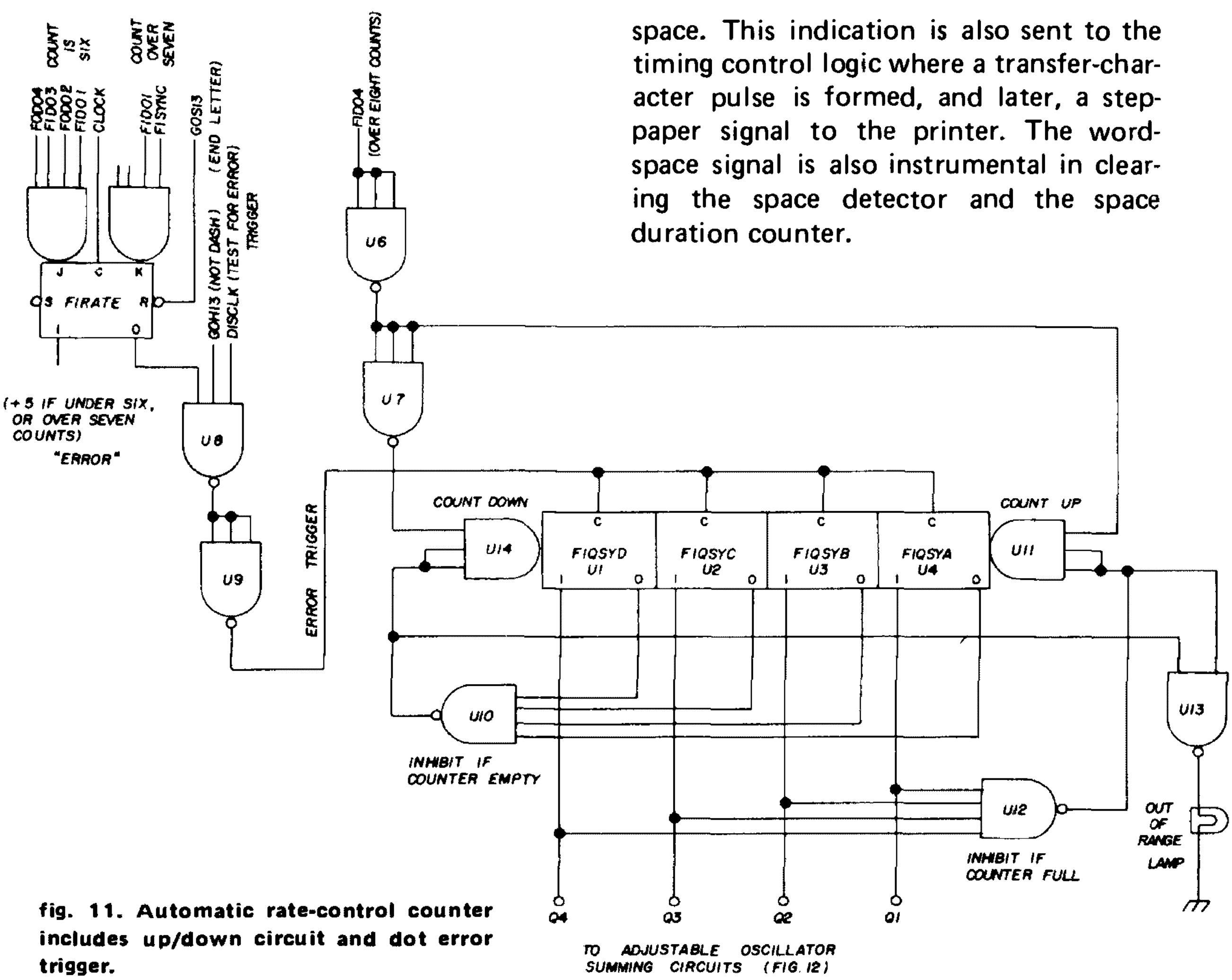


fig. 11. Automatic rate-control counter includes up/down circuit and dot error trigger.

of JK flip-flops. It counts clock pulses during the period that the *space* flip-flop is in a set state. This counter is reset either at the beginning of the next *mark*, or upon the detection of a word-space. Note that whenever a *sync mark* signal occurs it inhibits the input gates of the first stage (S1) of the counter. At the same time the clear-space-count signal is formed to clear the counter.

The character space decoder (gate G4 and inverter N2) detects the occurrence of a count of 13 in the counter. At this time the output of inverter N2 goes positive to indicate a character space. This indication is sent to the timing control logic where a transfer-character pulse is formed, and later, a print pulse.

The word-space decoder (gate G5) detects the occurrence of a count of 31 in the counter. The output of gate G5 goes negative at this time to indicate a word-space. This indication is also sent to the timing control logic where a transfer-character pulse is formed, and later, a step-paper signal to the printer. The word-space signal is also instrumental in clearing the space detector and the space duration counter.

The master oscillator and automatic frequency control circuit in **fig. 12** provides the clock signal for the AFF. This circuit was designed to maintain the *mark* count of 6 as the speed of the operator changed to track the sending operator.

The outputs from rate-control flip-flops F1QSYA, B, C and D in **fig. 11** drive the summing gate transistors Q1, Q2, Q3 and Q4. As these transistors turn on (or off) resistor R16 (82k), R17 (39k), R18 (20k) and R19 (10k) are connected (or disconnected) across shunt resistor R20. Resistors R5, R6 and R20 form a network that controls the current into the base of the current-charging transistor Q5. Unijunction transistor Q6, current generator Q5 and capacitor C1 form a relaxation oscillator which controls the basic frequency of the clock.

The rate at which the voltage across capacitor C1 rises is a function of the

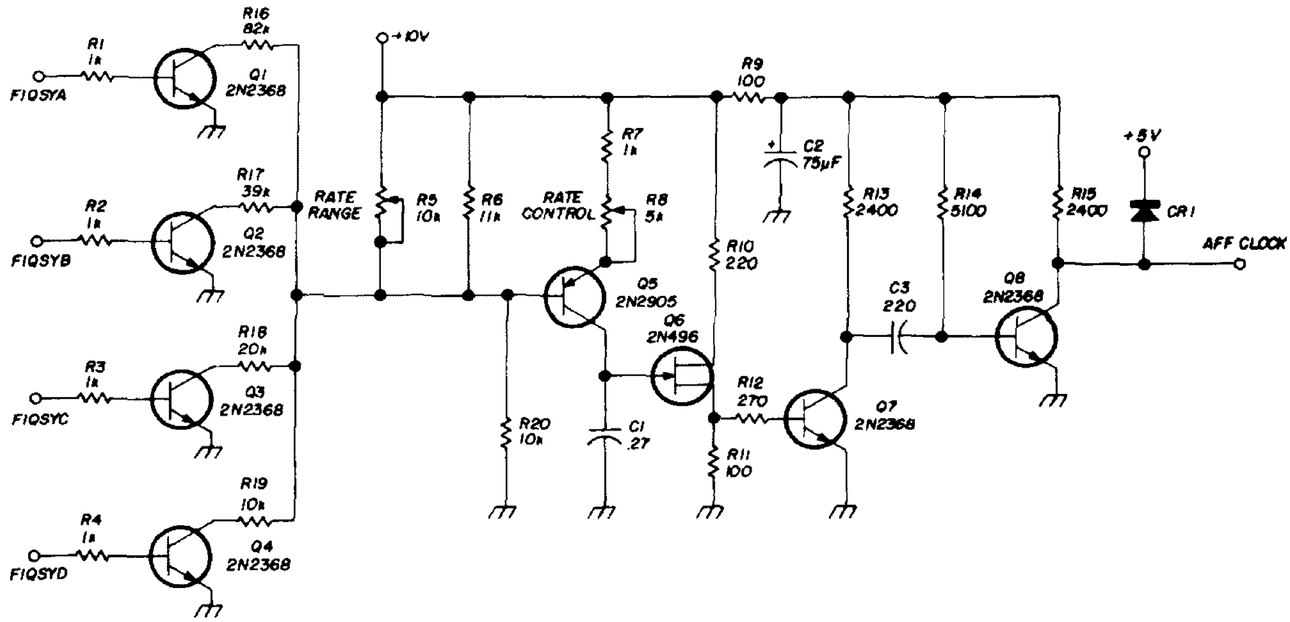


fig. 12. Adjustable oscillator and automatic frequency control circuit. Inputs to Q1, Q2, Q3 and Q4 are from *automatic rate-control counter* in fig. 11.

current generated by Q5 and the time required until Q6 fires at about 6.5 volts. This master clock signal controls the pulse-shaping single-shot multivibrator (Q7 and Q8) which generate a one microsecond clock pulse.

As the *mark* duration counter counts the clock pulses a feedback control signal is sent to the rate-control flip-flops (fig. 11) and updates the stored count if the dot length is more or less than the ideal six count; if it is six counts no change is made.

The value in this counter controls the summing network current and the clock frequency so a count of six is maintained in the *mark* generator. Since the control circuits in the automatic rate control counter will not respond to a dash signal (count over 13) the wpm rate changes only in relation to the average dot length.

translator

The heart of the automatic fist follower lies in its Morse/ASCII translator; this circuit is nothing more than a diode matrix network which compares the desired Morse code character with that of

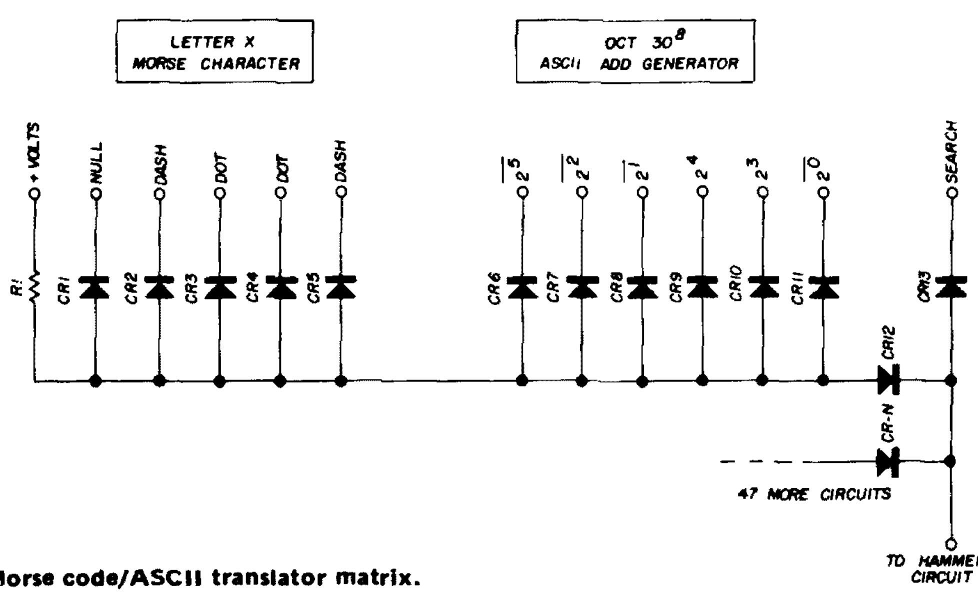


fig. 13. Morse code/ASCII translator matrix.

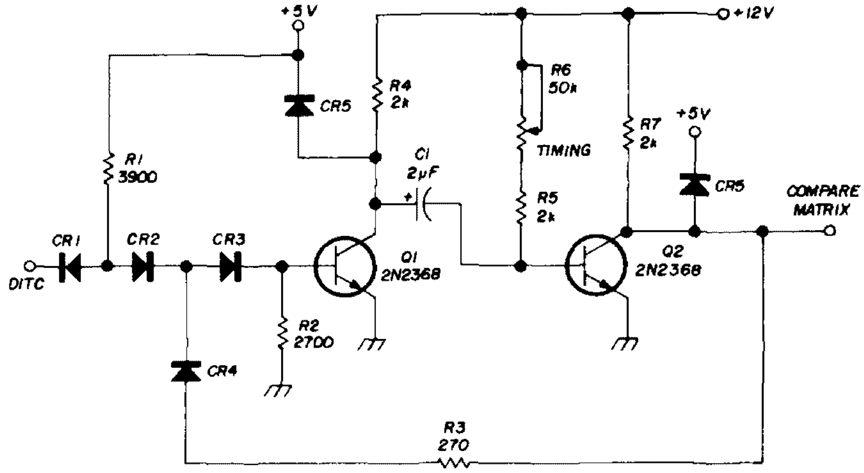


fig. 14. Print single shot is adjustable from 3 to 80 milliseconds; set to 40 milliseconds.

the corresponding letter on a print wheel. The position of any given character on this print wheel is detected by comparing, in a diode matrix, the value of the octal address generator with that of the desired character.

The operation of *one* compare line will be described with the aid of **fig. 13**. The eleven diodes in **fig. 13** are connected in a logic AND circuit. Five are connected into the Morse code character detect circuit; six diodes are connected to the ASCII address generator. Forty-eight such circuits are connected as OR logic circuits to AND with CR13 to generate the compare signal.

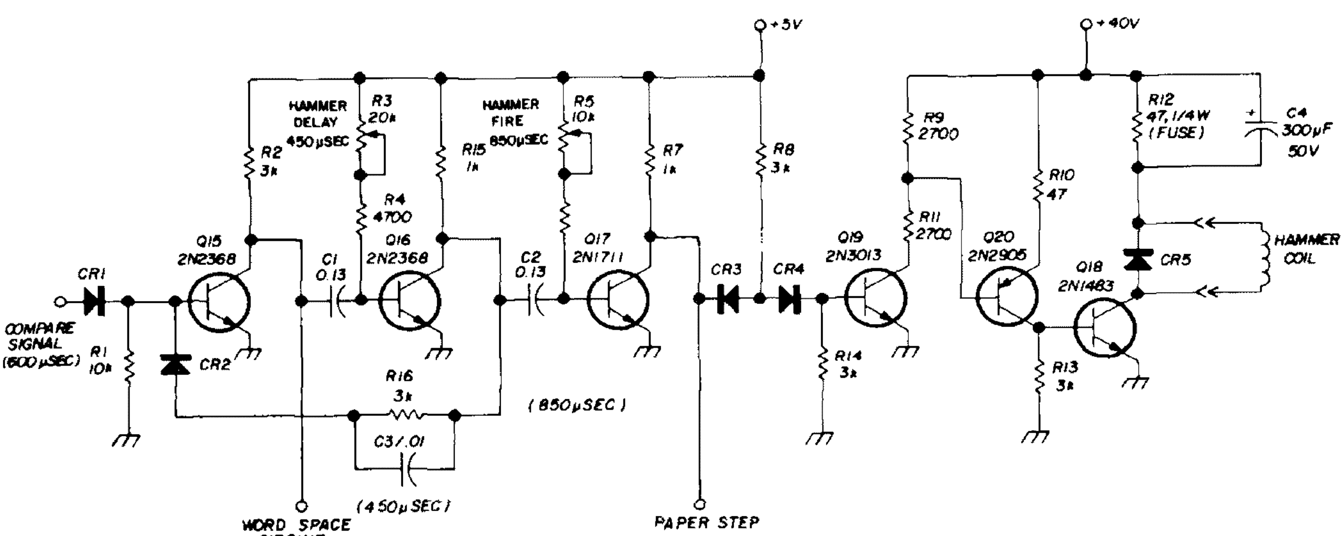
This compare signal (the result of all diodes in one line being at +5 volts) starts the hammer drive circuit. Only one of 64 combinations will be true at any one time for those diodes (CR6 through CR11) connected to the ASCII code generator. The generator is nothing more than six flip-flops connected to count from zero to 63 and repeat.

As the ASCII generator counts, it searches for those diodes which are at +5 volts. Since all diodes are at +5 volts at compare time, the compare line goes to +5 volts (through R1) and starts the hammer circuit. The received and decoded Morse code character comes in to diodes CR1 through CR5. As each new character is decoded by the AFF it is printed out through this translator.

print single-shot

The print single-shot (**fig. 14**) is a two-stage transistor circuit whose timing is enabled with a positive input trigger pulse applied to diode CR1. The positive pulse, from the Morse code translator

fig. 15. Hammer delay circuit. All diodes are 1N914s except CR5 which is a 1N4000.



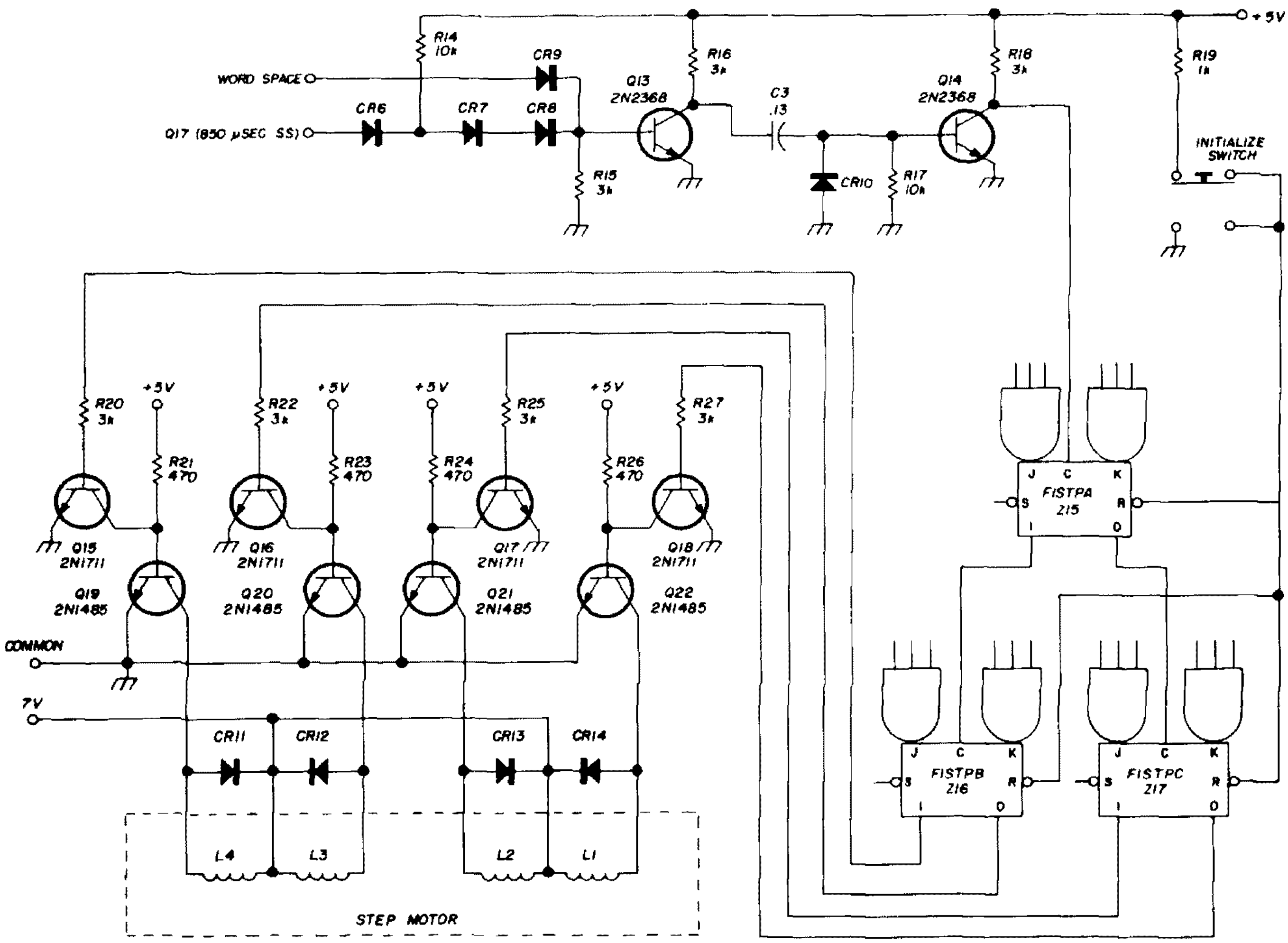


fig. 16. Paper-step control. Diodes CR11 through CR14 are type 1N4000. Integrated circuits Z15, Z16, and Z17 are Sylvania type SF200-08. Diodes CR6 through CR10 are 1N914s.

(character and word-space detector) logic, triggers transistor Q1 on. With Q1's collector clamped to ground, capacitor C1 starts to discharge through resistors R5 and R6. The rate of discharge is determined by their time constant. With the base of transistor Q2 below ground, a positive voltage is applied from the collector of Q2 through R3, CR3 and CR4 to the base of transistor Q1, keeping Q1 turned on until the charge across C1 decreases to a level that allows transistor Q2 to turn on.

This action occurs each time a character is formed. R6, the timing resistor, is set to give a 40-millisecond pulse output from Q2. This drives the Morse code to ASCII diode compare matrix. It is during this time that the print wheel will complete one revolution, allowing the desired Morse character to be synchronized with the comparing ASCII character on the print wheel.

print control

Control of the mechanical functions within the printing mechanism requires very precise timing signals. Three single-shot circuits were developed for controlling this basic timing. Single-shot circuits of standard design include two or more stages for each timing function; this circuit (fig. 15) uses five stages to generate the three separate timing signals.

The first section of the timing circuit allows time for the print wheel to synchronize the desired character (embossed on the face of the rotating print wheel) with the print hammer impact interval. The compare signal from the Morse code/ASCII translator signals the approach of the desired character toward the hammer impact point.

A positive pulse at diode CR1 turns on transistor Q15; this clamps one end of capacitor C1 to ground and starts the

timing cycle for R3, R4 and C1. The cycle ends when C1 charges and allows transistor Q16 to return to the on state in approximately 450 microseconds.

The end of that timing cycle causes C2 stepping motor and control the direction of rotation, one step for each input trigger. Transistors Q15, Q16, Q17, Q18 and Q19 through Q22 serve only as current amplifiers. Fig. 17 shows a func-

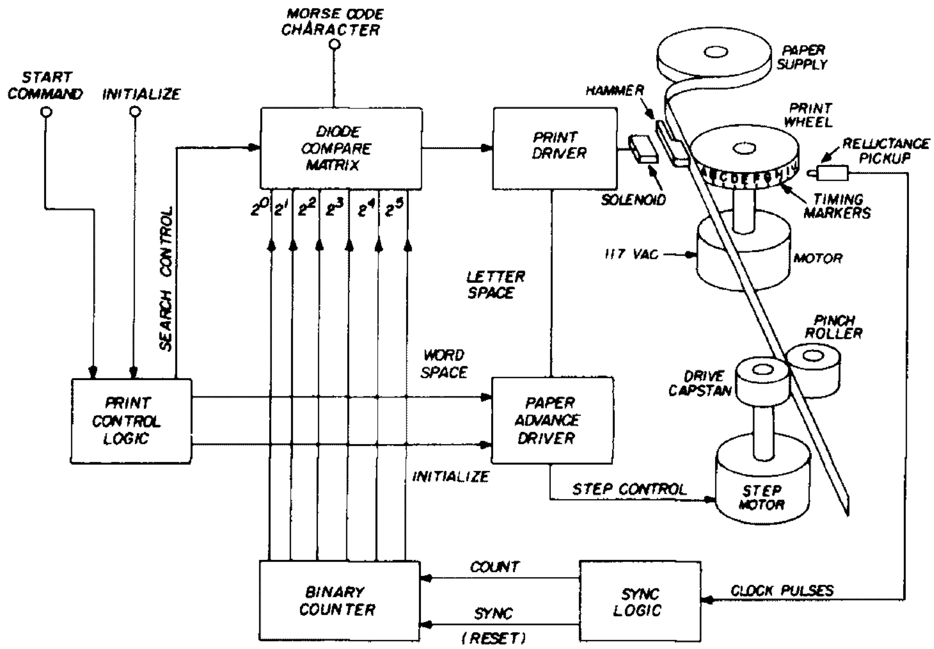


fig. 17. Block diagram of the basic strip printer.

to be clamped to ground by the collector of Q16, which starts the hammer *on* drive signal. The output pulse from Q17 drives not only transistor Q19, but also starts the delay time for the stepping-motor trigger.

At the time when Q17 was positive, transistor Q13 was biased on, and capacitor C3 was discharging through diode CR10. Upon the termination of the print hammer pulse transistor Q13 is switched off, allowing R16 to drive current into the base of transistor Q14. Transistor Q14 is normally biased off by R17 holding its base to ground. When Q14 turns on, a fast negative-going pulse is generated to trigger the JK flip-flop Z15 in the step-motor control circuit which advances the paper tape.

The step motor is controlled by three JK flip-flops, Z15, Z16, Z17 (see fig. 16). The four control states are: 1010, 1001, 0101 and 0110. These control states enable the field currents within the tional block diagram of the basic strip printer and control.

construction

The photograph shows an overall view of the four circuit breadboards with the strip printer in the foreground. Nearly any method of construction can be used as long as the interconnecting wires are not over one foot long. This is because the TTL logic chips are very fast and noise would trigger them.

The first board has all of the AFF logic and those transistors and components which make up the master oscillator and automatic rate control.

The second board holds the auxiliary lamp display. This is not needed for operation of the AFF logic and was used as a display until the printout control was designed.

The third board holds the electronic circuits required by the printer. The fourth board contains the diode compare

circuits. Not much can be said about the matrix diode board except that it's awkward, takes a lot of room, and we wish it wasn't needed.

During construction, there is one point that you should be aware of: use care in the ground and voltage supply lines. They must be short. Number 24 awg was used on the prototype with bypass capacitors (1.5 μF disc) every six inches or so.

adjustment

Operation of your receiver with the AFF is not unlike its normal operation in the CW mode. The only special consideration is that the audio beat note produced by the desired CW signal must fall within the bandpass of the 800-Hz active filter.

With the prototype unit tests showed that with 200-Hz bandwidth tuning on a Drake TR-4 was nice and smooth, allowed good QRM rejection yet received CW speeds up to 100 wpm without dropping data due to frequency rolloff.

The only other requirement is that the receiver's output level be at least 0.2 volts rms and above the noise level to trigger the active filter.

Operation of the AFF's logic is perhaps the easiest of all. You have only to tune in a CW signal, set the *run/calibration* switch to *calibrate,* adjust the rate control until the dot lamp starts to flash, then turn the switch back to *run.*

This technique should have you receiving automatic Morse code with the drudgeries left to the AFF machine. You can sit back and read the copy.

Even with the automatic tracking ability of the machine there is always some guy with a lead foot — a lid who can do things with Morse code even old man Morse did not know about. You don't have to give up; simply go to manual operation (calibrate position) and ride the rate control till the copy is clear. The dot lamp may not always light, due to the idiosyncrasies of this type of sending, but the copy should be readable.

We put a sign over our machine which reads, "This machine cannot think! It can copy Morse code."

The results of several months of on-the-air operation with the AFF were very enlightening. Machine-sent Morse code, even at speeds exceeding 100 wpm, was duck soup for the AFF. Surprisingly, even quite poor hand-sent Morse code was copied almost solidly due to the automatic tracking feature of the unit.

The major cause of errors was due to signal fading below the noise level or extremely strong interference right on top of the signal. This pointed up the extreme importance of the shaper, filter and keyer sections in the front end. More time was spent arriving at the optimum design for this area than any other portion. The bandpass was narrowed as much as possible while still allowing fairly high speed code to pass through the filter.

Limiting the maximum code speed to about 60 wpm is reasonable for amateur use and allows a sharp narrowband filter. However, the prototype AFF has an adjustable bandpass circuit that allows reception of very high speed code when desired.

The AFF was built with the fairly expensive high-speed transistor-transistor-logic (T^2L) integrated circuits because they were readily available. However, these are not necessary; the lower cost, low-speed ICs are ideal for this application since they are less susceptible to noise and less critical in operation. The same logic functions are available in these low-cost circuits, so substitution is quite straight forward.

The AFF prototype performs very well. It is a simple step from here to a low-cost version. In addition, the output interface can be adapted to a teleprinter or visual display. The age of ICs and automation is truly with us and it is starting to have considerable impact on amateur communications.

references

1. P. Horowitz, W2QYW, "Perfect Code at Your Fingertips," *QST,* August, 1965, page 11.
2. Selfridge and Neisser, "Pattern Recognition," *Scientific American,* August, 1960.
3. P. Laakmann, WB6IOM, "Signal Detection and Communication in the Presence of White Noise," *ham radio,* February, 1969, page 16.

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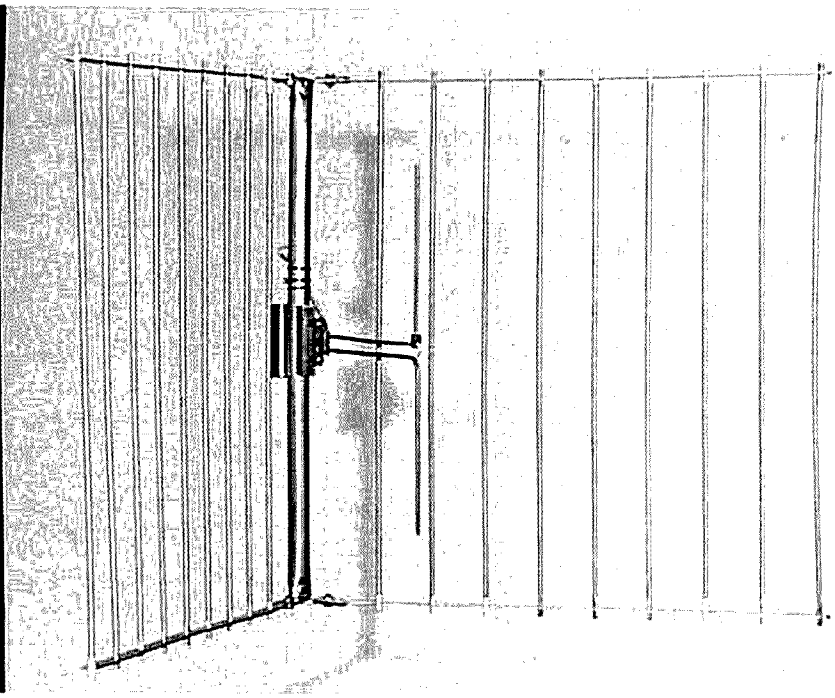


432-MHz corner-reflector antenna

This miniature 7-dB antenna may be used by apartment dwellers, or as a reference for antenna measuring

Fred Telewski, WA2FSQ

Here is a small 432-MHz antenna which will suit the needs of apartment dwelling amateurs or those who want a standard gain reference antenna. This 7-dB antenna makes a nice portable window unit as well as an excellent comparision antenna for larger arrays.

Antenna contests have shown that many complex arrays (both amateur and commercial) do not perform as expected. This is due in part to the fact that many gain measurements are made via E- and H-plane pattern integration that assume a lossless structure. This is often not true because the structures are lossy. In addition, directivity is a prerequisite for gain but gain is not required for directivity. Since the corner reflector described here is simple and foolproof it will provide a good reference for large arrays.

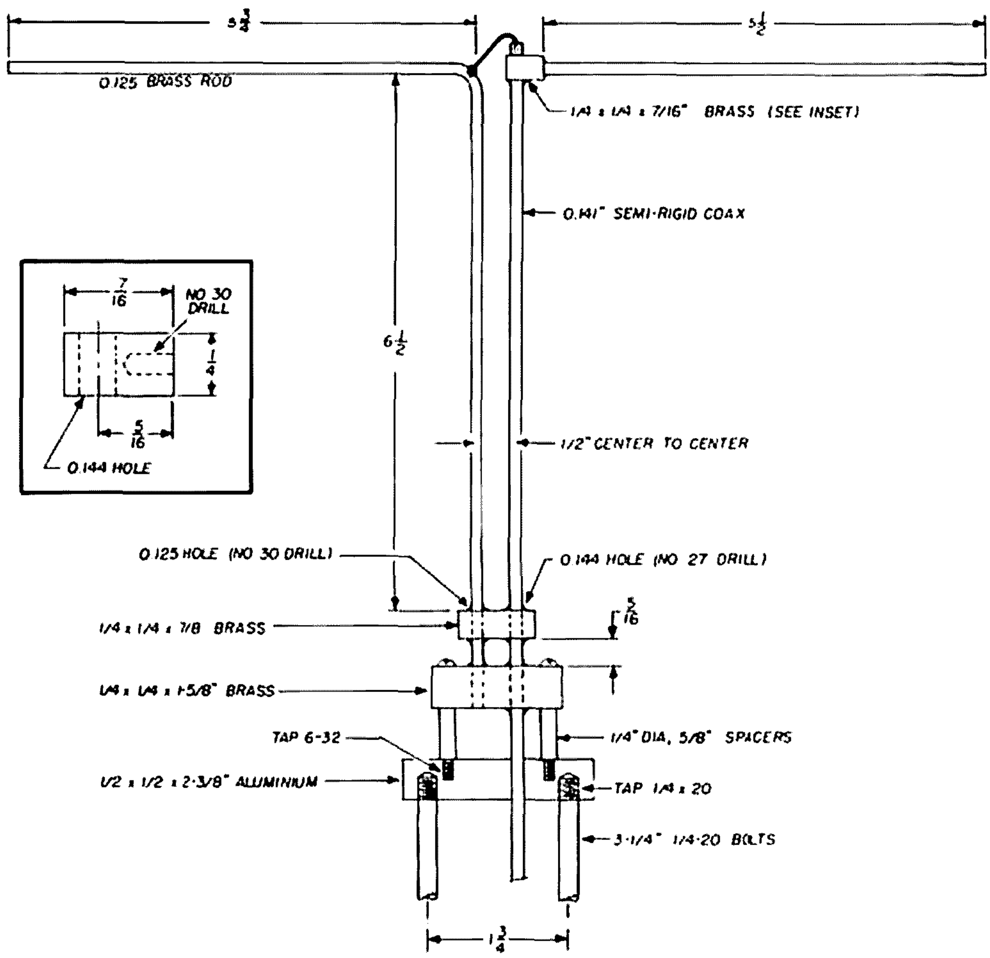
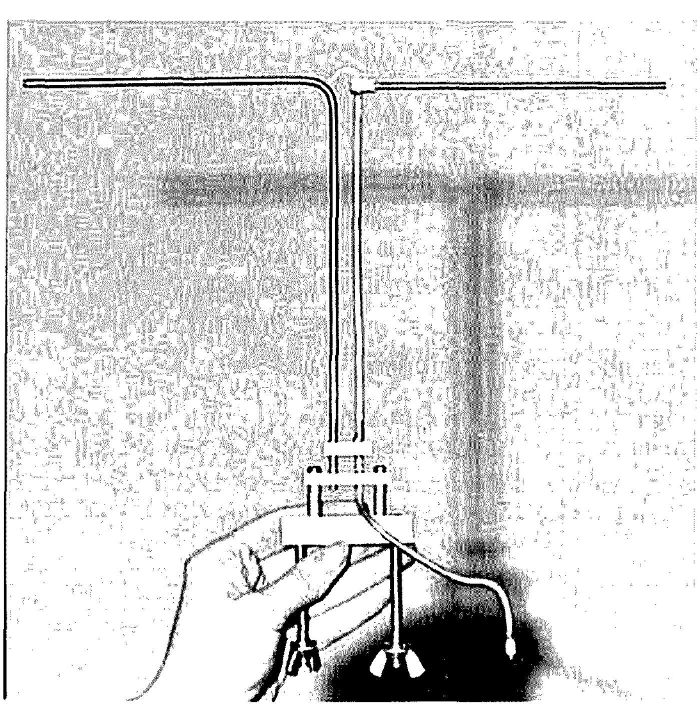


fig. 1. Construction details for the driven dipole element of the 432-MHz corner reflector antenna. Reflector is built from commercial uhf-tv antenna.

the reflector

The corner reflector is a modified JFD uhf television antenna which is priced at around $5.00. Start modifying the antenna by removing the U-bolt and saddle clamps; drill out the rivet which holds the bow-tie assembly in place. Discard the bow-tie and replace it with the driven dipole described in fig. 1. The original bow-tie will not function properly at 432 MHz.



Driven dipole assembly.

For a completely collapsible antenna, drill out the spring-loaded rivets which hold the reflector panels to the back bar and replace them with 8-32 screws and wing nuts.

driven dipole

The driven dipole consists of a coaxial line (Phelps-Dodge 0.141-inch semirigid line*), a balancer and a dipole. A variety of connectors for the semirigid line are available from Phelps Dodge Corporation;

*Phelps Dodge Communications Corporation, 60 Dodge Avenue, North Haven, Connecticut 06473.

the one recommended here is a BNC female type PDM 952-001. If you don't want to purchase a special connector, you can make one out of a BNC type UG-88/U connector by making a bushing which will adapt 0.141-inch line to the parts carefully with fine sandpaper or steel wool and apply flux before soldering. Do not use a torch; the excessive heat may cause the Teflon to expand and burst the outer conductor of the semi-rigid line.

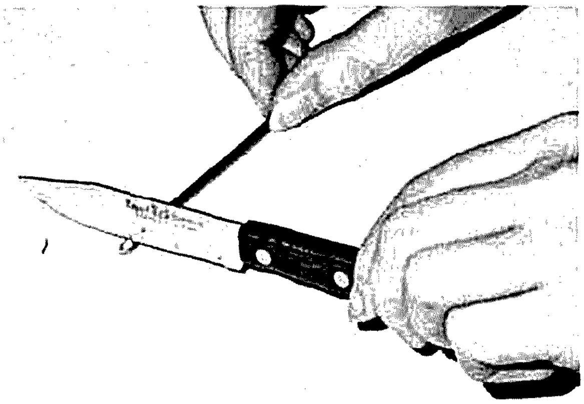


fig. 2. First step in preparing semi-rigid coaxial line is to score the outer conductor with a sharp knife.

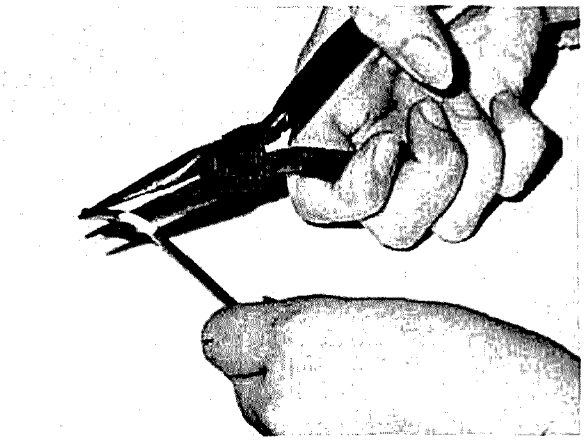
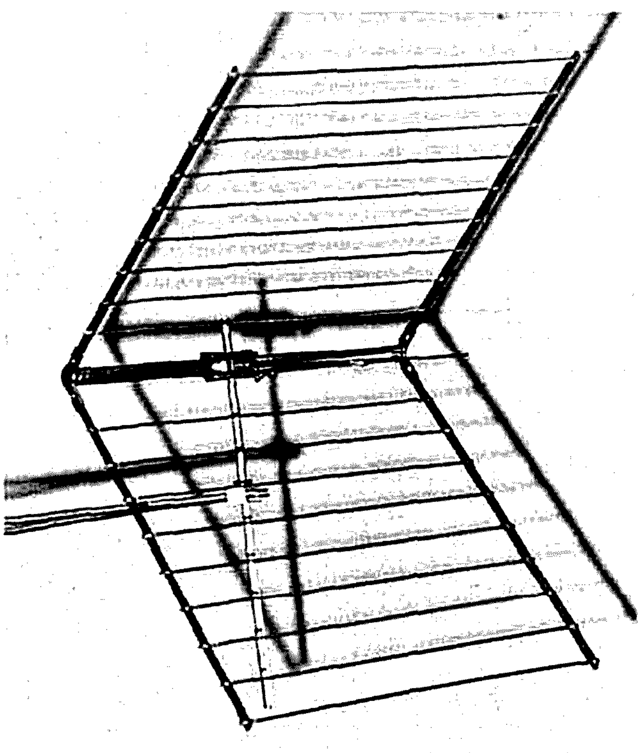


fig. 3. Break the outer conductor by grasping the end with pliers and rocking it back and forth.

rear nut and soldering it in place.

The driven dipole parts should be soldered with a 250-watt iron. Clean the



Corner reflector attached to window-mounting bracket.

A few hints are in order for working with the semirigid coaxial line. This cable may be prepared by scoring the outer conductor with a sharp knife (see fig. 2). When the score is complete, grasp the end lightly with a pair of pliers and rock from side to side (see fig. 3). This separates the outer conductor from the Teflon dielectric. Now the outer conductor may be removed. Carefully score the Teflon (do not nick the center conductor) and remove it to expose the center conductor (see fig. 4).

performance

This antenna will deliver approximately 7 dB gain over a dipole. This design should yield a vswr of 1.2:1 or less if the dipole is carefully constructed. The 0.141-inch semirigid line will handle 200 watts of rf output power quite easily. For higher powers (500 to 600 watts) use 0.250-inch semirigid coaxial line.

radiation hazard

The potential hazards of rf radiation should always be considered when working in close proximity to vhf and uhf

antennas. In general, it is good practice not to stand directly in front of any vhf-uhf antenna excited with 10 watts or more of rf. At these frequencies the rf energy causes thermal heating of body tissue which may have undesirable effects on body organs, notably the eyes. Since rf radiation effects are dependent on the average power dissipated by the tissue, a-m and fm are potentially worse than CW or ssb. This is due to the continuous high-level carrier associated with a-m and fm.

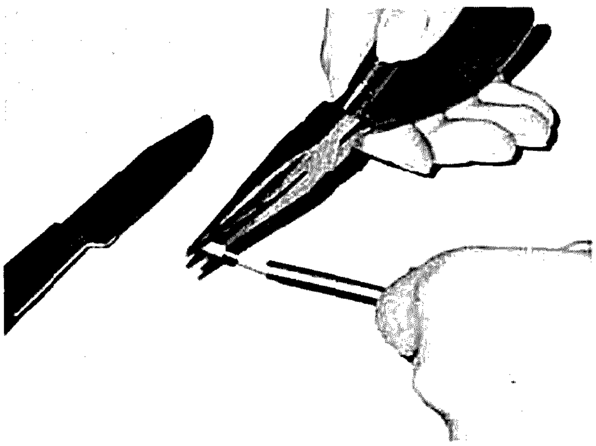


fig. 4. Score the Teflon dielectric and remove. Be careful not to nick the center conductor.

With 100 watts of rf delivered to this antenna, the field strength 6 feet in front of the antenna is well below the 10 mW/cm^2 safety standard commonly accepted in America. Radiation off the sides and back of the antenna is less than 0.1 mW/cm^2. (These measurements were conducted with a calibrated power meter and a Narda microwave radiation monitor, model B86B3.)

The antenna may be mounted out a window and swung from side to side without any significant rf field concentration in the building. The window brace shown in fig. 5 is useful for mounting the antenna in an apartment window.

reference antenna

Amateurs who want to use this antenna as a reference might wonder why you would use a gain antenna as a reference instead of a dipole. This is because of the serious errors often introduced with a dipole reference because the dipole is prone to picking up reflections from the back, as well as ground reflections from the front.

In many cases the comparison antenna can be made to look like a champ or a dud by simply moving the reference dipole two or three feet in one direction or another so that reflections are arriving in or out of phase. Therefore, a reference antenna with some gain and a single major lobe is desirable. It eliminates back-lobe contributions so the major reflections which effect the antenna are the forward ground type; the effects of these may be determined by moving the antenna in the vertical plane.

In general, measurements with respect to this type of reference antenna are a good deal more consistent than those made with respect to a dipole. Since the amateur is not usually concerned with the absolute gain of his system but rather whether he has made an improvement or not, consistency in the measurement technique is most desirable.

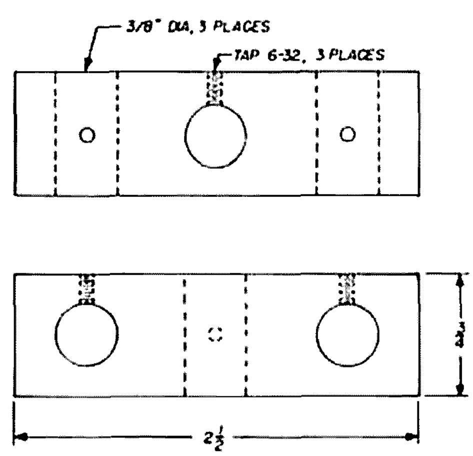
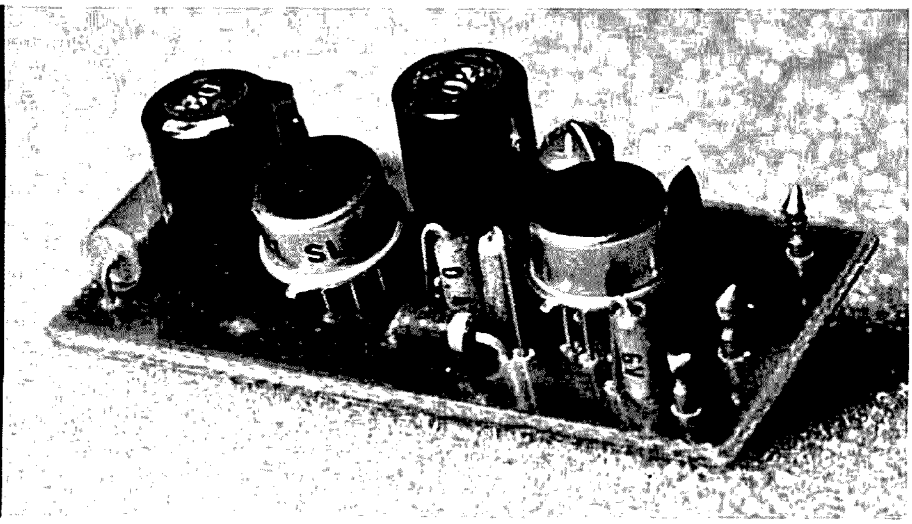


fig. 5. Window mounting blocks are made from 3/4" aluminum bar stock. Mounting rods are 3/8" aluminum tubing, 2-feet long. Complete installation requires 2 mounting blocks and 3 rods.

In conclusion I would like to extend my thanks to Ted Miller of Phelps Dodge for his cooperation. I would also like to acknowledge R. Knadle, K2RIW, for his help and comments.

ham radio



miniature microphone preamplifier with agc

An integrated-circuit speech preamplifier that features nearly constant output with large variations in speech input

James M. Bryant, 29 Arle Gardens, Cheltenham, England

For many applications it is convenient to be able to reduce the dynamic range of an audio signal. The modulation depth of a transmitter is best kept quite high; a constant value of modulating signal is useful to prevent overmodulation. Also, if speech is being taped at low tape speeds too low a signal will result in a poor signal-to-noise ratio; too large a signal will produce distortion.

This article describes a speech amplifier system with agc which will provide an rms output stable within 2 dB for nearly 40 dB input range. The circuit uses two integrated circuits and only ten discrete components, and occupies a printed circuit measuring only 22 x 47 mm (0.8 x 1.8 inch) on which there is also room for a miniature microphone.

An ideal speech agc system must adapt quickly to an input level and follow a

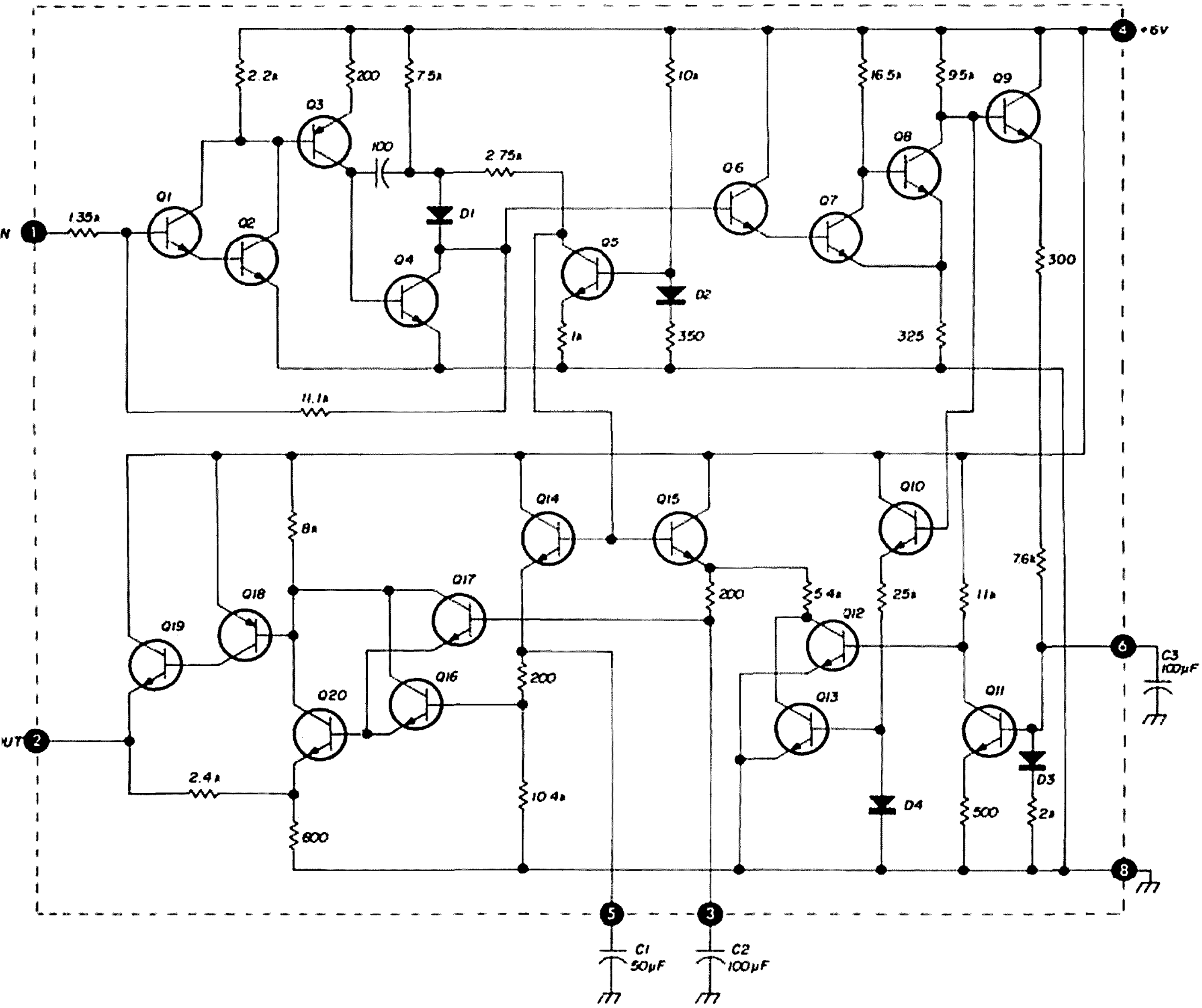
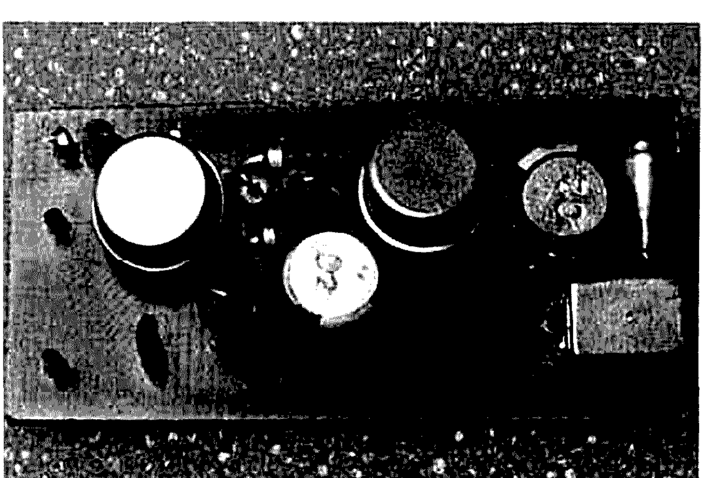


fig. 1. Internal circuit diagram of the Plessey SL620C agc generator.

fading signal but *remember* its level if speech suddenly stops as it does in pauses between words. However, if the pause becomes too long the system must quickly revert to full gain.



agc generator

The Plessey SL620C agc generator* meets the requirements detailed above and will also produce short pulses of agc to suppress short noise bursts while remembering the agc level at which it was operating when the noise burst first arrived. A circuit diagram of the SL620C is shown in fig. 1; its response to a varying audio signal is plotted in fig. 2. Capacitors C1, C2 and C3 are connected externally.

Circuit operation is as follows: the incoming signal is amplified by an af amplifier using transistors Q1 to Q4; its

*Plessey integrated circuits may be purchased from Plessey Electronics Corporation, 170 Finn Court, Farmingdale, Long Island, New York 11735.

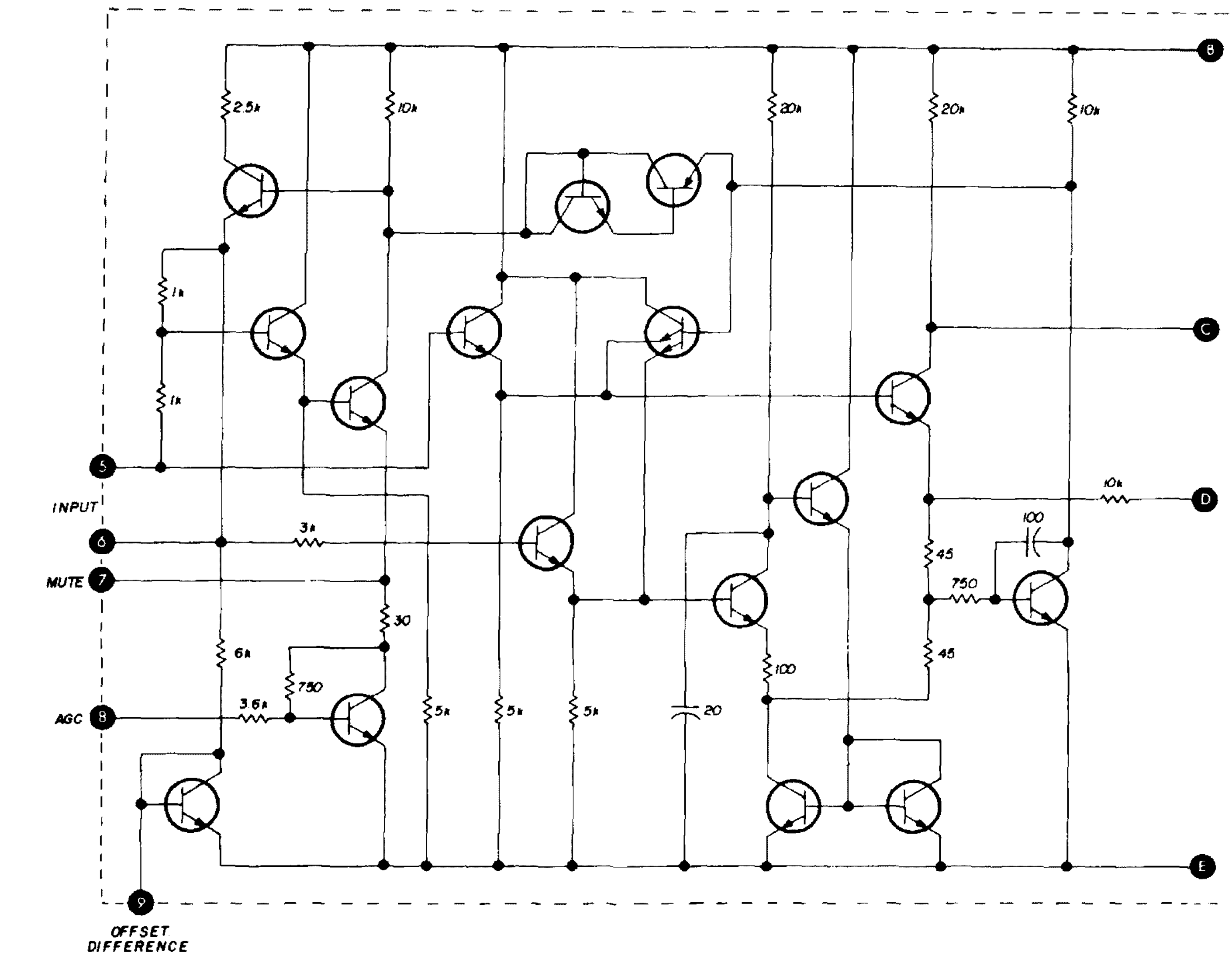


fig. 3. Circuit diagram of the Plessey SL630C audio amplifier.

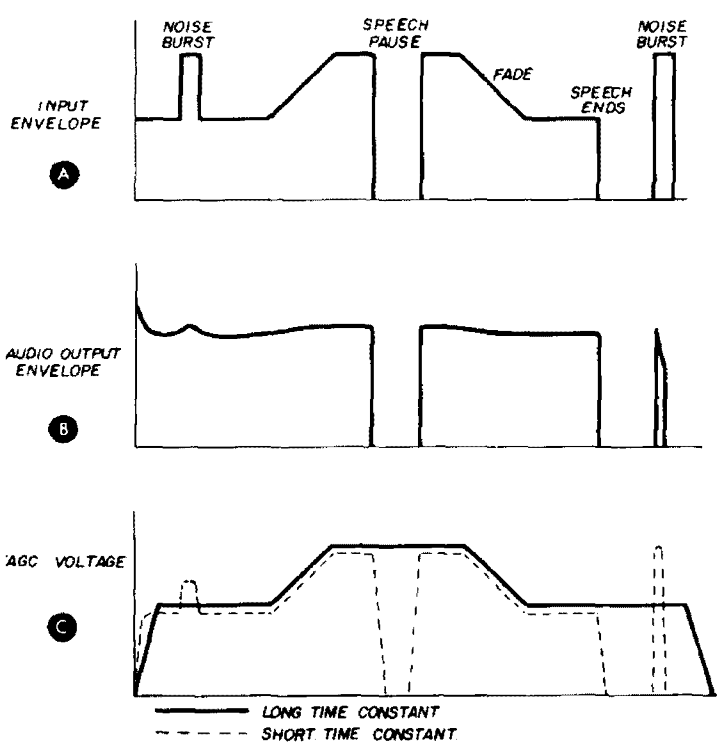
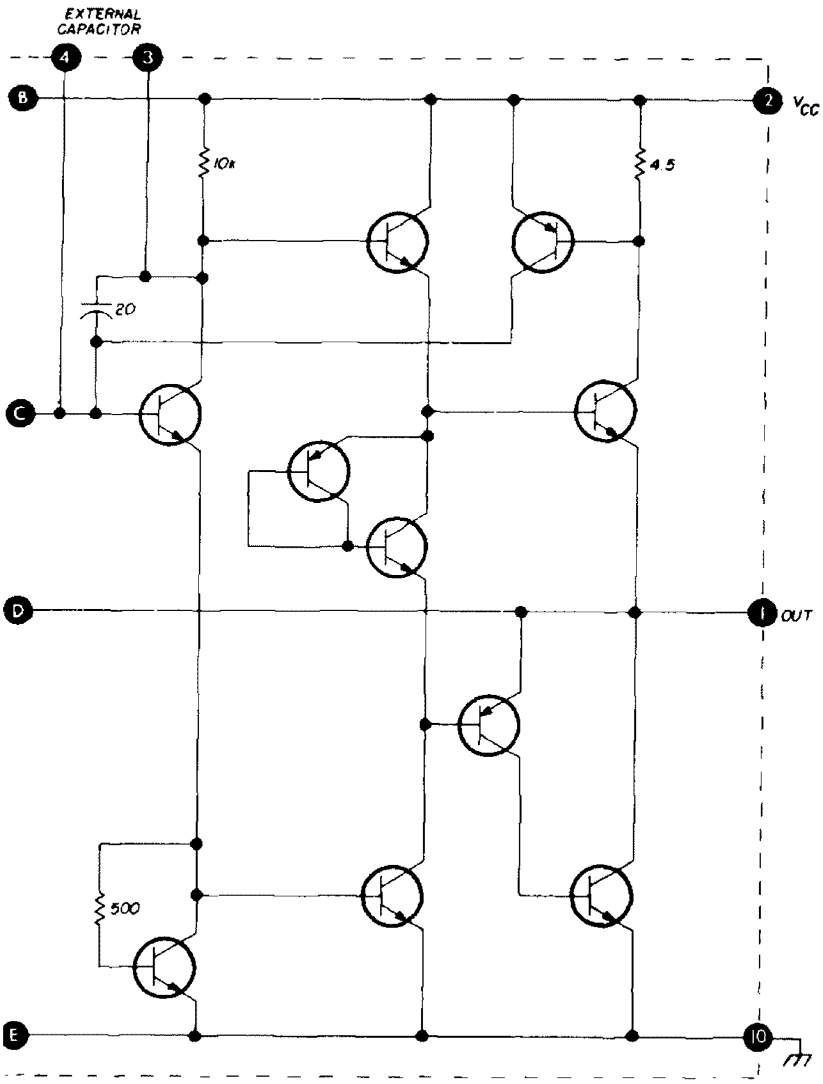


fig. 2. Response curves for the SL620C.

dc level is shifted by Q5 and applied to two detectors, Q14 and Q15. These detectors charge C2 and C2 respectively. Capacitor C1 has a short charge and discharge time constant; C2 has a longer one. Both of these capacitors drive a dc output amplifier (Q16 to Q20); this dc output amplifier is connected so that whichever capacitor has the larger potential controls the output.

If both C1 and C2 have the same charge, a slight offset ensures that C2 will control the output. Thus, the incoming signal will quickly establish an agc output via Q14 and more slowly establish a charge across C2 via Q15.

If sufficient signals are present at the input to the SL620C the trigger circuit (transistors Q6 to Q8) will operate to provide a fast discharge path for C2 via Q10 and Q13. This allows the input signal



to fade up to 20 dB per second and still be tracked by the agc. However, if the input signals are too small the trigger stops operating and C2 is not discharged. This enables the agc signal to *remember* its original level during pauses in speech.

After a long pause capacitor C3, which is charged by the trigger circuit via Q9, will discharge and allow Q12 to conduct, rapidly discharging C2 and allowing the agc signal to drop, thus restoring full gain to the controlled amplifier. If a burst of noise occurs the short time-constant system consisting of Q14 and C1 will briefly control the system to reduce its level.

audio amplifier

The Plessey SL630C audio amplifier circuit is illustrated in fig. **3.** In addition to its agc connection (pin 8) it has a mute terminal (pin 7) and a reference (pin 9) which is used in conjunction with a linear potentiometer to provide logarithmic volume control when manual gain control is required. This is shown in fig. **4.** The amplifier is muted when the mute

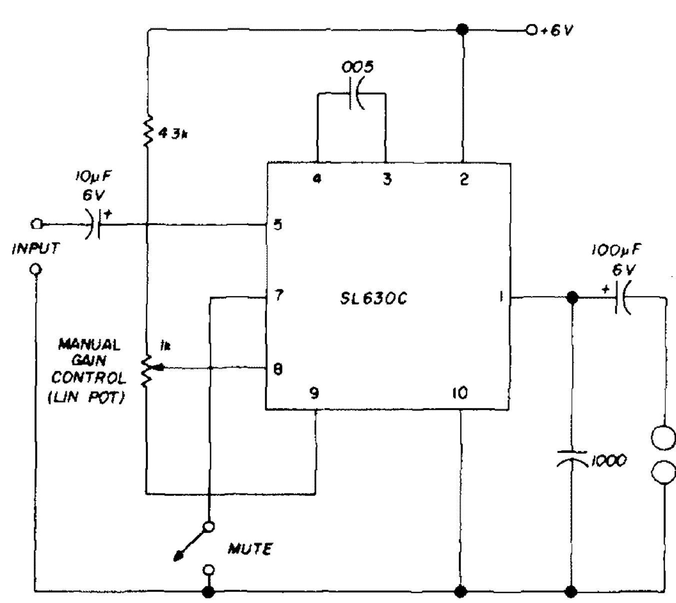
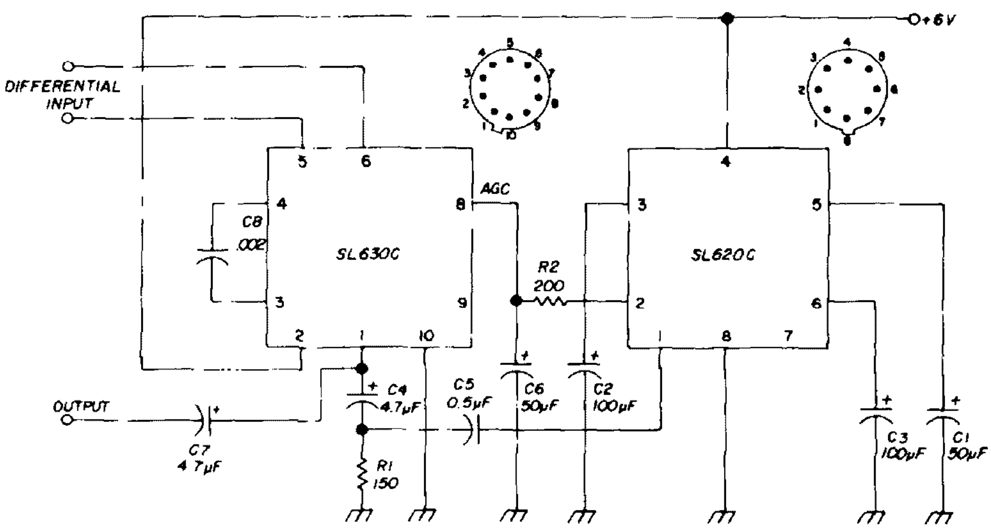


fig. 4. Practical circuit for using the SL630C as an amplifier with manual gain control.

terminal is grounded. With mute control two or more SL630C devices may be connected to a common output without extra loading as long as all but one are muted at any one time.

The SL630C audio amplifier has an internal 20-pF capacitor which provides 6-dB per octave rolloff above 800 kHz. However, in normal operation this capacitor is shunted by an external generator. If this happens low-frequency feedback will occur and low-frequency oscillation or motorboating will take place. This will not occur if the indicated component values are used for C4, C5 and R1.

The agc amplifier is built on the miniature printed-circuit board shown in **fig. 6**. To use this board the smallest possible components must be used. When



C1,C6	47-µF, 6-volt, subminiature tantalum	C5	0.47-µF, 35-volt, subminiature tantalum
C2,C3	100-µF, 6-volt, electrolytic	C8	0.0022-µF, subminiature ceramic
C4,C7	4.7-µF, 10-volt, subminiature tantalum	R1	150-ohm, 1/8-watt
		R2	220-ohm, 1/8-watt

fig. 5. Practical speech preamplifier circuit features wide dynamic range. Output is 80 mV rms.

capacitor (connected between pins 3 and 4) to provide a lower frequency 3-dB point at a frequency defined by the formula

$$C = \frac{10^9}{2\pi f}$$

where C is in pF and f is the frequency in Hz.

agc amplifier

The complete audio agc amplifier is shown in **fig. 5**. Since the SL630C has some gain between the agc input point and the output, care must be taken so low-frequency signals from the SL630C do not reach the input of the SL620C agc preparing the printed-circuit board particular care must be used because the conductor widths and separations are very small.

Except for the power supply and output connections, all holes in the printed-circuit board should be 1/32 inch. The holes for the power supply and output terminals should be the correct size for the terminals you use. All soldering should be done as quickly as possible with a small soldering iron and fine rosin-core solder. Before you install the ICs, pin 7 of the SL620C and pins 7 and 9 of the SL630C should be cut off as close to the can as possible. These pins are not used.

Before connecting the agc speech amplifier to a power supply double check that all devices are properly connected and have the correct polarity. Also check to see that there are no solder bridges between the conductors of the circuit board. The power supply should be 6 volts ±0.5 volt. If you use a battery connect a 500-μF decoupling capacitor across it.

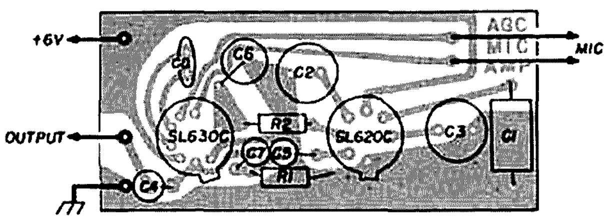


fig. 6. Full-size layout of miniature printed-circuit board for the speech preamplifier.

operation

The finished agc amplifier system should deliver about 80 mV rms for a wide range of inputs. The microphone I used, a microminiature Knowles Electronics type 1590 which measures only 4 x 6 x 8 mm, is mounted on the circuit board. The complete unit is small enough to mount easily in a small microphone case. Microphones with output impedances below about 5k ohms and output of a few millivolts or more may be used with this preamplifier.

If single-ended input is required, it should be connected between ground and pin 5 of the SL630C through a capacitor not larger than 1 μF. In this case pin 6 should be left open. However, this arrangement is not recommended because instablity can result.

To test the agc preamplifier use an ac voltmeter (vtvm or oscilloscope) and an audio oscillator to check that output remains constant as input is varied. Otherwise, you can connect a pair of headphones to the output and alternately shout and whisper into the microphone while an assistant in another room listens for changes in volume.

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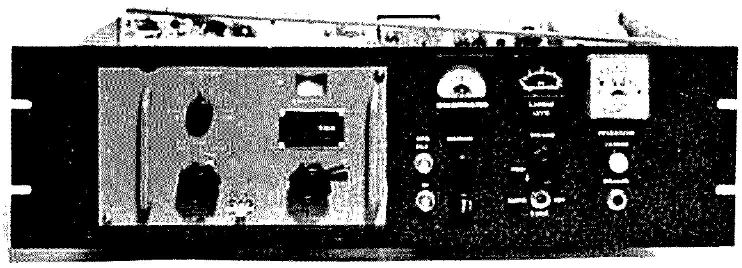

tunable vhf fm receiver

This tunable vhf fm receiver is based on the use of surplus building blocks

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The tunable fm receiver described in this article provides good vhf fm reception by simply adding a suitable converter. The basic tuning range of the receiver is 8 to 16 MHz. Therefore, the overall tuning range with a 2-meter converter is 142 to 150 MHz, and with a 6-meter converter, 48 to 56 MHz. With the addition of a vhf converter, the 8- to 16-MHz tuning range becomes the first i-f; the second i-f is at 455 kHz.

The sensitivity of the receiver is 0.6 microvolt for 20-dB quieting (first i-f) and 0.35 microvolt for 20-dB quieting (with the 2-meter converter). Image rejection is more than 70 dB (with the 2-meter converter).

Some of the optional features of this tunable vhf fm receiver are carrier-operated relay (COR), deviation meter, relative carrier meter, crystal control and automatic frequency control (afc).

construction

The tunable fm receiver is fairly easy to put together since it consists of commercially built and tested units as shown in fig. 1. As I am interested in 2-meter fm, I use an Ameco CN-144 converter ahead of the tuning unit.

The 8- to 16-MHz tuning unit is a tuning drawer from a surplus AN/FRR-49 series receiver made by TMC. The original receiver was also known as the AN/FRR-502. Several different tuning drawers are available for this receiver; a list of the

more useful units for amateur use is given in **table 1**. These are available on the surplus market. In addition, a large number were released to Navy MARS a few years ago.

The tuning drawers consist of two tunable rf amplifiers, a mixer and oscillator. A reactance tube is included in the oscillator circuit for afc. The oscillator may also be crystal controlled. These features, including a crystal socket, are available on the front panel of the tuning drawer.

In the original receiver the main FRR-49 chassis contained the power supply, 455-kHz i-f amplifier and audio circuits. The main chassis is known as the FRR-2 or R-5007/FRR-502. If you have patience you can find mating connectors for the tuning drawers on the surplus market; if you're in a hurry they can be purchased from an industrial electronics distributor – Cannon part number DPD-1204-34P.

The i-f strip used in the tunable fm receiver described here was originally part of a Motorola LO4AB uhf base-station receiver. The actual i-f strip I used is a Motorola uhf type TA-183 designed for

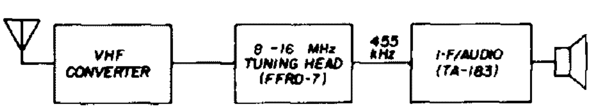


fig. 1. Block diagram of the tunable vhf fm receiver.

service in the 404- to 420-MHz range. For my use I removed the rf sections, mixer, first and second i-f stages and local-oscillator chain from the chassis. This left only the audio, squelch and third i-f at 455-kHz. Since the circuit included a 455-kHz bandpass filter this was included in my finished receiver.

A few items were added during assembly including a zero-center microammeter to allow comparison of stations on channel. Another microammeter is used to indicate relative carrier strength, and a deviation meter is helpful in setting up fm

table 1. FFR tuning units most suitable for amateur use.

frequency range	tuning unit	alternate part no.
2.0 — 4.0 MHz	FFRD-5	TN-5010/FRR-502
4.0 — 8.0 MHz	FFRD-6	TN-5011/FRR-502
8.0 — 16.0 MHz	FFRD-7	TN-5012/FRR-502
16.0 — 32.0 MHz	FFRD-8	TN-5014/FRR-502

transmitters. A carrier-operated relay is used to operate a signal lamp or key a tape recorder for logging; this circuit is the standard Motorola type.

The relative carrier meter simply reads the first-limiter grid current. The deviation meter consists of an amplifier/buffer

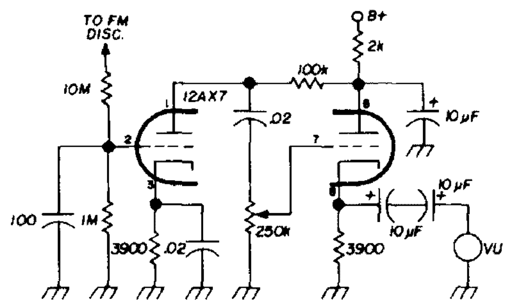


fig. 2. Deviation meter for the tunable fm receiver.

and meter driver shown in **fig. 2**. Discriminator response was plotted by using a signal generator, frequency counter and vtvm. A 1-kHz tone was then inserted at an appropriate level (taken from the response curve) and the meter trim potentiometer adjusted for proper meter indication.

Although I didn't use the afc provision it is available if you want to use it. In some cases you may find that the correction voltage from the discriminator has the reverse sense. In this event reverse the connections to the plates of the dis-

*Complete schematic diagrams of the FFRD-7 tuning head, Motorola TA-183 i-f/audio system and power supply are available for $.25 from *ham radio* magazine, Greenville, New Hampshire 03048.

criminator tube or use an IC operational amplifier as an inverter.

The power supply for the tunable fm receiver was designed around an old tv power transformer. The B+ supplies for the various receiver sections were tailored to the requirements. The i-f strip requires 200 volts at 70 mA and 6.3 Vac at 2 amps; the surplus tuning drawer requires 250 volts at about 20 mA, 150 volts regulated at 30 mA and 6.3 Vac at 1.4 amp. The Ameco two-meter converter needs 100 to 125 volts at 25 mA and 6.3 Vac at 0.9 amp.

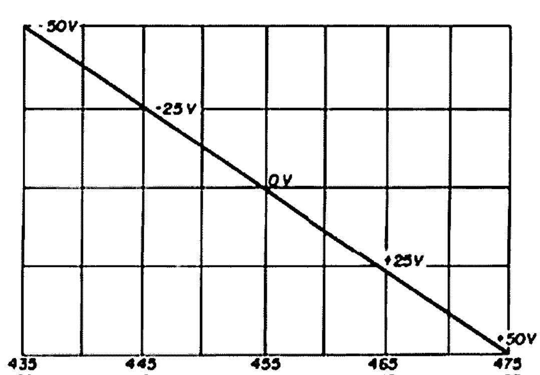
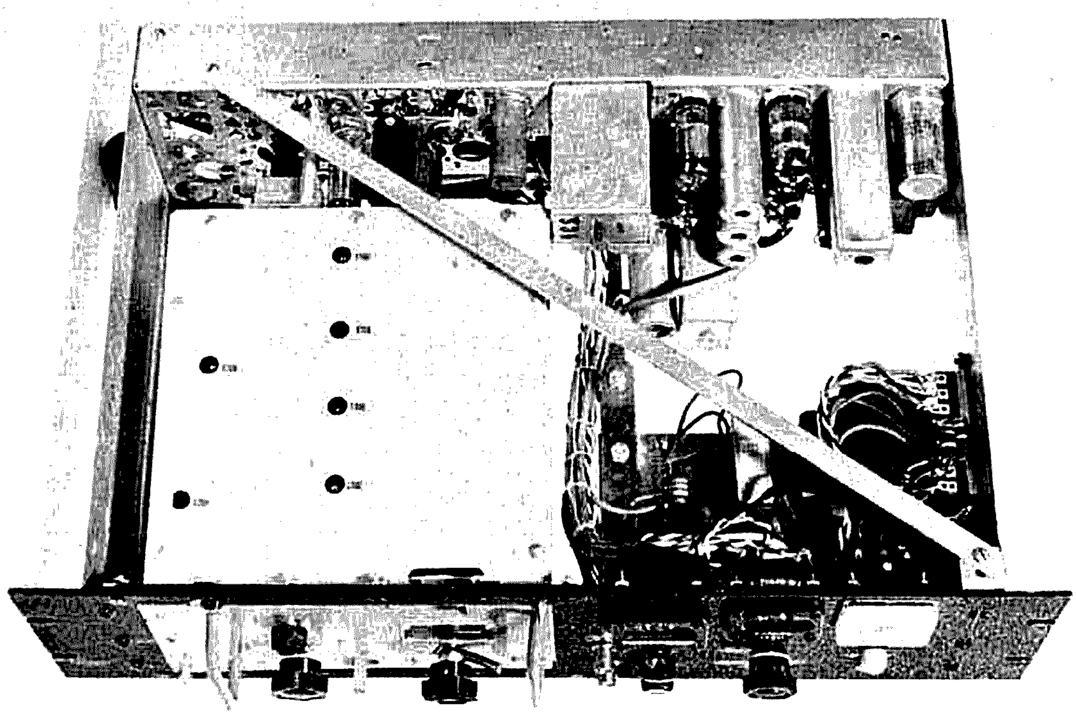


fig. 3. Discriminator response of the Motorola TA-183 i-f strip.

vhf converters

Since 2 meters is my primary interest, I used an Ameco CN-144 converter ahead of the 8- to 16-MHz tuning drawer. The converter was set up for output at 10 to 14 MHz. The rf gain control on the converter was adjusted to the point where first-limiter grid current just started to increase: overall sensitivity of the receiver was 0.35 microvolt for 20-dB quieting at 147-MHz input.



Top view of tunable vhf fm receiver shows surplus tuning drawer to left. Modified Motorola TA-183 i-f strip is at rear. Power supply is next to the front panel on the right.

summary

There are many variations to the basic receiver discussed here. It is especially nice to be able to use existing items from around the station to build a useful, quality performer as described here. I have found this tunable receiver to be very useful in setting up my station fm equipment and netting it to frequencies used in my home area. In addition, I can monitor any channel I want without disturbing my regular fm equipment or spending the money for an additional crystal.

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six-meter conversion of the Heathkit SB200 linear amplifier

The SB200 can be converted into a high-efficiency linear for 50-MHz in less than two hours

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Most 50-MHz conversions of high-frequency rf power amplifiers involve modifications to the existing tank circuit. In most cases the number of turns on the 10-meter coil is reduced so the circuit will tune in the 50-MHz range. The 6-meter Heath SB200 conversion described here uses a tank circuit that is designed specifically for 6 meters. This provides maximum plate-tank circuit efficiency through the use of optimum L/C ratios, constant load to the exciter and maximum tuning ease.

design parameters

This six-meter conversion was based on two essential ground rules: nothing was to be done that would change the front panel, and the unit must be left in such a condition after the conversion that it could be easily restored to its original condition. Both of these aims have been realized.

Only three items must be purchased for this conversion: two pi-network tuning capacitors and 16-inches of copper tubing. The complete conversion can be completed in about two hours and requires no special tools.

step-by-step conversion

1. Remove the amplifier from the enclosure and remove the top shield cover.

2. Remove the two T160/572B power tubes and store in a safe place.

3. Unsolder both parasitic suppressors (PC1 and PC2) from the 1000 pF blocking capacitor (C24). Remove one turn from each suppressor and place aside. Unsolder the lead from plate choke RFC1 to C24.

4. Unsolder and carefully remove L7 (10-, 15- and 20-meter coil). Be careful not to break the switch connections on wafer B of the bandswitch.

5. Unsolder the connections from L6 (80- and 40-meter coil) to wafer A and B of the bandswitch. L6 may be left in the rf compartment if you wish. I chose to remove it.

6. Unsolder C26 (100 pF) from wafer A of the bandswitch and dismantle from C25.

7. Unsolder the heavy bus bar connections which connect both halves of C28 (loading capacitor). Remove the

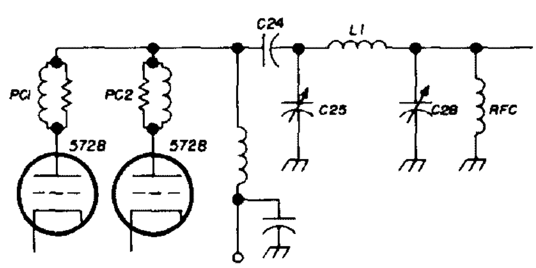
heavy bus wire connection which comes from wafer B on the top side of the chassis to C28 on the underside of the chassis (marked X in **fig. 2**).

8. Remove RFC3 (1.1 mH). Remove the RG-58/U cable between the relay and C28.

9. Unscrew all mounting hardware from C28 loading capacitor. To remove this capacitor from the chassis, it is necessary to loosen the hardware holding the front panel; following removal of C28 secure all panel hardware.

10. Remove the mounting hardware from C25 plate tuning capacitor and remove it from the chassis.

11. Install the new E. F. Johnson **154-1** air variable (C28). Use the front mounting hole of the original loading capacitor as a guide and drill and tap a 6-32 hole to match the mounting bracket of the new loading capacitor. Install the capacitor using star washers on all hardware.



C25	9-38 pF air variable (E. F. Johnson 154-11)
C28	12-244 pF air variable (E. F. Johnson 154-1)
L1	4 turns 3/16" copper tubing, 1¼" diameter, spaced 3/16" between turns (see fig. 3)
RFC	Ohmite Z-50
PC1,PC2	parasitic suppressors

fig. 1. Modified plate circuit for the 6-meter SB200.

12. Select the new E. F. Johnson 154-11 capacitor (C25), and using the front mounting hole of the original capacitor, drill a 6-32 hole to match the bracket of the 154-11. Install the capacitor.

13. Install the tuning knobs and tighten all front-panel hardware. Align the knob markings with the left edge of the *tuning* indices with both capacitors fully meshed.

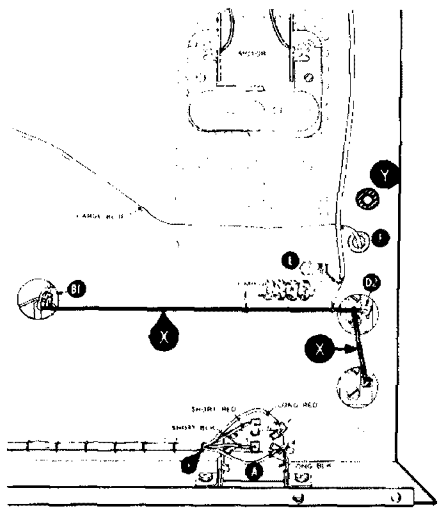


fig. 2. Under-chassis modifications to the SB200. Remove wires marked with X. Drill new 3/8" hole at position Y.

14. Install the plate coil and the 1000 pF coupling capacitor as shown in fig. **4**. Use a 6-32 x ¾-inch threaded brass or aluminum standoff between the tuning capacitor and the 1000 pF coupling capacitor. Be sure to use star washers on all mechanical connections.

15. Install a 6-32 solder lug on the 1000-pF capacitor. Connect the 1000-pF capacitor to RFC1 with solid number-**12** copper wire.

16. Install parasitic suppressors PC1 and PC2 and plate cap assemblies to RFC1.

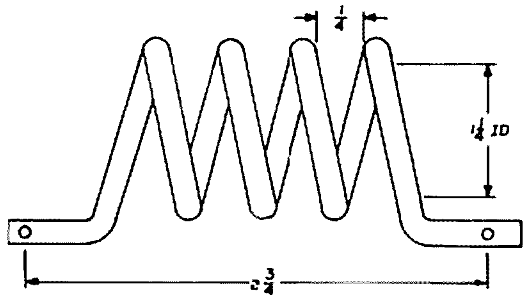


fig. 3. Construction of the plate-tank coil for the 6-meter SB200. Material is 3/16" copper tubing. Two small holes at ends are drilled for 6-32 screws.

17. Install a 6-32 solder lug on the rear stator plate of C28 – nearest the shield wall. Use a star washer.

20. Drill a 3/8-inch hole in the chassis (location Y in fig. 2) just above grommet F. Install a ½-inch rubber grommet.

21. Cut the RG-8/U exactly 9½-inches long. Strip 2½-inches of the unstripped end.

22. Unravel the braid and twist securely. Trim to a length of 1½ inches.

23. Using the 2½-inch stripped end of the RG-8/4 cable, attach and solder the braid to solder lug K (fig. 2).

24. Strip ½ inch of insulation from the inner conductor of the cable and solder it to lug 7 of the relay.

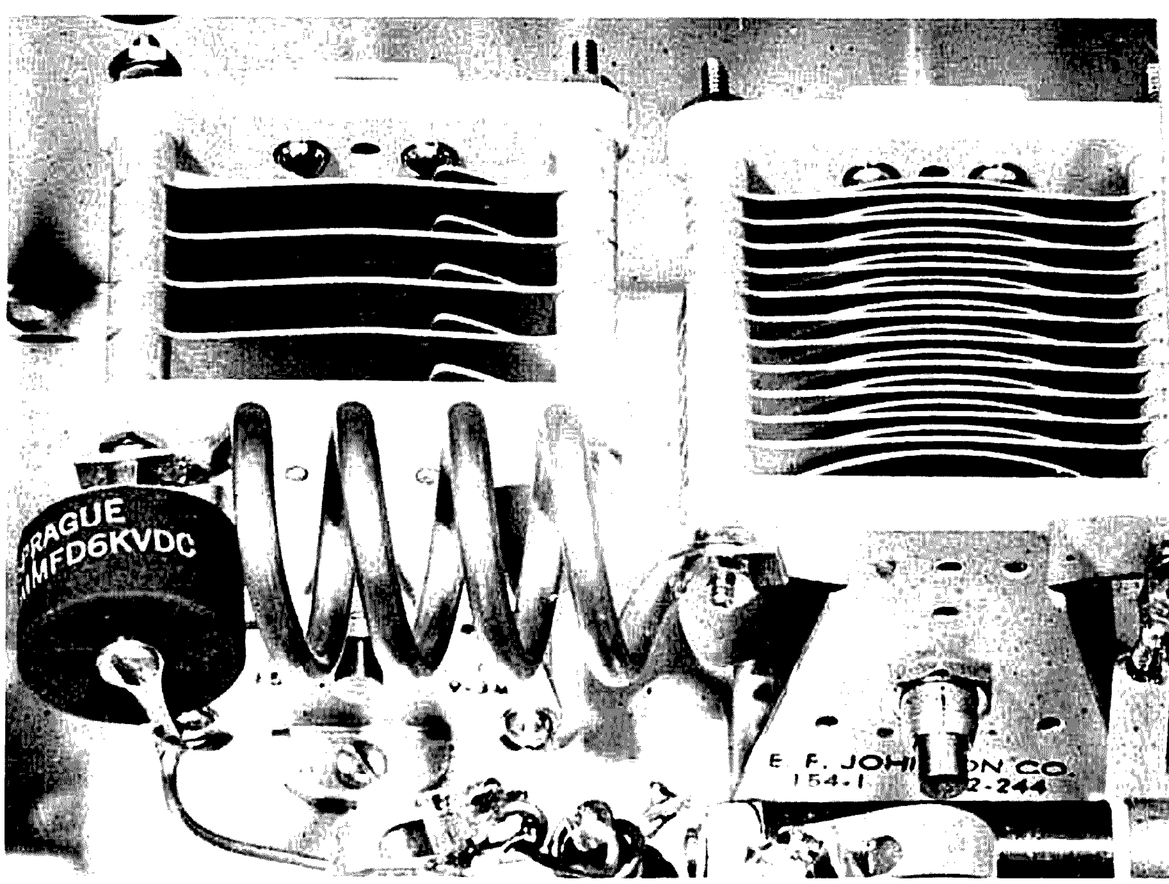


fig. 4. The plate-tank coil is mounted between solder lugs mounted on the pi-network capacitors. The 1000-pF coupling capacitor is mounted on a threaded metal standoff.

18. Strip 6 inches from one end of a 20-inch length of RG-8/U.

19. Unravel the shield braid and cut the braid 1½-inches long, measured from the end of the vinyl jacket. Twist and tin the braid.

25. Solder the shield braid at the other end of this piece of RG-8/U to solder lug E (Fig. 2).

26. Strip ½ inch of insulation from the inner conductor and push this end of the cable through the grommet at Y.

27. Solder the inner conductor to the empty solder lug on loading capacitor C28.

28. *Install an Ohmite Z-50 RF choke* between the solder lug in step 27 and the ground solder lug located on the rear mounting bolt of loading capacitor C28.

This completes the modifications in the rf compartment. Install the T160/572B power tubes and attach the plate caps. The only steps remaining are those necessary to modify the input circuit. Refer to fig. 5.

1. Disconnect and remove the 68 pF capacitor from coil CA to the ground lug at AE.

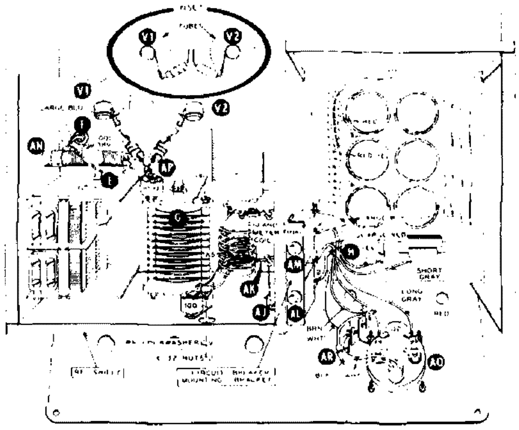


fig. 5. Above-chassis construction of the SB200 before modification.

2. Connect a 15 pF capacitor between lug 2 or CA to the ground lug at AE.

alignment and tune up

1. Connect the amplifier to the exciter; connect the keying line.

2. Set the bandswitch to the ten-meter position (this is now the permanent six meter position).

3. Set the load capacitor knob to 2 on the dial.

4. Set the plate tuning capacitor at half mesh.

5. Turn on the exciter and the linear amplifier. Allow three minutes for both to warm up.

6. Place the meter switch of the amplifier to the plate current position.

7. Key the exciter and apply enough drive to show 300 mA on the meter.

8. Adjust the tune (plate) control for a dip.

9. Increase exciter drive and adjust the amplifier for a dip at 500 mA plate current (use both plate tuning and loading controls). On this amplifier the dip occurs with the plate tune control at 40% mesh and the load control at 2½ on the dial.

10. Change the meter switch to the *rf out* position and adjust the *Sensitivity* control for half-scale indication.

11. Adjust the plate tuning control slightly for a maximum rf output indication.

Note: During all tuning and adjusting do not allow a key down condition for more than 30 seconds at a time.

12. Key the exciter and apply drive for a plate meter indication of 200 mA.

13. Adjust coil CA for maximum plate current indication.

This completes the 6-meter modifications to the SB200. This linear can be driven with any exciter which has an output of at least 35 watts. The unit in the photographs belongs to WA1LVF; Roy is driving it with a Heathkit SB110A. Tests conducted with Bird Thruline wattmeters on both exciter and linear, loaded into a Heath Cantenna, showed 60 watts output from the SB110A and linear amplifier output of 600 watts.

No instability has been experienced with this conversion. If you should exper-

ience instability, a stub neutralizing rod located between the power tubes, mounted on a ceramic feedthrough and connected to the bifilar choke with a 2-turn link should stabilize the amplifier.

optional modification

The small motor and fan used to circulate air around the power tubes leaves much to be desired for my tastes; I prefer lots of air around power tubes. The miniature Pamotor muffin fan can be purchased from most supply houses for about $10. In some areas it may be available as surplus.

Remove the original fan and fan motor. Use a nibbler tool or a fine coping saw to cut out a square hole to fit the outside dimensions of the Pamotor fan. Be sure to leave enough metal on the diagonal corners to accommodate the mounting ears. Mount the fan on the underside of the chassis; don't worry, there is just 3/16-inch of clearance between the fan and the bottom edge of the chassis.

Mount a 2-lug terminal as shown in fig. 6. Connect the black wires which went to the original fan motor to the tie points and the Pamotor fan leads to the same tie points. When the fan is installed in the SB200 cabinet there is more than enough room from the bottom of the fan to the cabinet. All SB-line enclosures are perforated, and this really helps air circulation. With the installation of the muffin fan air circulation has been improved by a factor of at least five. The fan motor is very quiet.

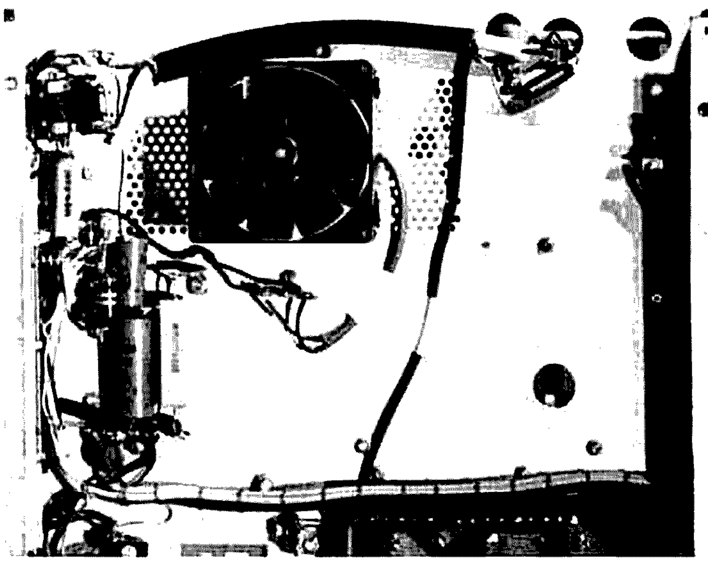
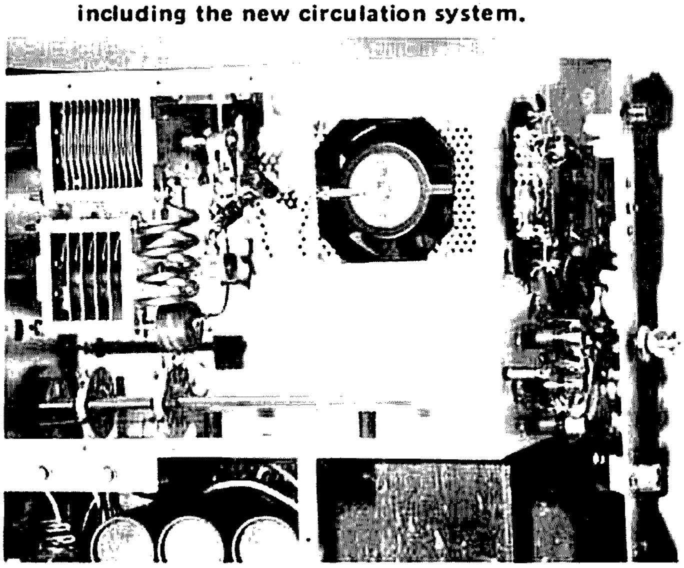


fig. 6. The Pamoter muffin fan is mounted directly below the plate area of the final tubes. The 2-lug tie point is mounted to the right of the fan motor.

Top view of the completed SB200 conversion including the new circulation system.



conclusion

The cost of the conversion is, to my mind, most reasonable. The two capacitors in the pi-network were ordered directly from E,F. Johnson for $15.63 including postage. The Pamotor fan was $9 locally and the copper tubing was 45c at the local hardware store. With the exception of the 15 pF capacitor, the new grommet and the length of RG-8/U cable, all other parts were salvaged from the original circuitry.

The SB200 has been a very popular linear amplifier and now, with the introduction of the SB220, they should be more plentiful on the used market. This conversion provides very respectable power output from a table-top linear which uses inexpensive tubes. I will be glad to answer inquiries if accompanied by a sase. My thanks to Roy Colella, WA1LVF, for furnishing the SB200 to experiment on and to Dick McGinn, WA1IMS, for the photographs.

ham radio

weak-signal reception in cw receivers

Suggestions for controlling receiver-generated noise, with some interesting test results

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Much has been written about weak-signal reception. It is well known that a cw contact can be maintained long after signals using other modulation modes have become too weak to resolve. The limiting factor in copying weak signals is usually the noise generated *within the* receiver. Noise from external sources is also a problem, but even this noise may be minimized by using the correct approach in receiver design. It is the purpose of this article to demonstrate methods of reducing internally generated noise so that really weak signals may be copied under "quiet" conditions.

receiver noise

One method of assessing the capability of a receiver to receive weak signals is to replace the antenna with a dummy load and adjust the audio-noise output to a measured value, say -10 dB, when the receiver is adjusted to maximum sensitivity. A signal is then applied that will increase the audio output by 10 dB. The injected-signal level, in microvolts, is then a measure of the limiting signal-to-noise ratio. A receiver with a 1-microvolt sensitivity under these conditions is considered average; a sensitivity of ¼ microvolt is excellent.

The noise at the receiver audio output under dummy-load conditions is caused in a tube-type receiver by the tubes. It's usually stated that most of the noise comes from the *first* tube in the receiver; one good reason for this statement is that the tube furthest from the audio has the most amplification behind it. Sensitivity, therefore, is to some extent a matter of rf amplification versus rf tube noise. Because two tubes make more noise than one, we need really high amplification from a single rf stage.

Receiver noise, finally, increases with amplification. This is why so little noise accompanies a strong signal. We reduce the gain; and the noise, being weaker, disappears first. It's when we are working at the limit of amplification that noise intrudes, and any reduction of noise is tantamount to increasing the level of the *incoming signal.*

There are various ways of reducing noise in relation to signal. One is to reduce the number of tubes — there is nothing quieter than a one-tube "blooper!" Another is to improve the rf stage sensitivity. A third way is to narrow receiver bandwidth.

Noise covers a wide frequency range. Any reduction in receiver audio-frequency bandwidth will reduce noise. For cw reception a very narrow frequency bandwidth is desired. If the audio amplifier is designed to attenuate all frequencies except that which responds to the received signal, then unwanted audio frequencies and attendant noise will be

reduced. Such reduction in bandwidth not only reduces much receiver-generated noise; external noise, such as that from power leaks and auto ignition, will also be reduced.

gain control

An important consideration in design of a receiver for weak-signal performance is that of gain control. An enormous amount of amplification is needed to increase a 0.1-microvolt input signal to a 1-milliwatt audio level. Without an effective gain-control system, receiver overload will occur when receiving very weak signals in the presence of powerful signals or local noise. A sensitive rf stage and its associated mixer, for example, will seldom overload in the presence of a powerful signal, so this part of the system requires neither manual nor automatic gain control (agc). Early i-f stages, on the other hand, are subject to overload in the presence of strong signals and require efficient gain control circuits.

Agc circuits have been developed for cw-signal reception, but they are more complex than those for phone signals. In a cw agc system, the agc voltage must be derived from a rectifier isolated from the beat oscillator, otherwise the rectified bfo voltage will reduce receiver gain even in the absence of incoming signals. Therefore, most amateurs use the audio-gain control; but this control has its limitations.

An audio-gain control is an input attenuator. The audio amplifier runs at full gain all the time, but its input is limited by the audio-gain control. When a strong signal is received, the product detector may be overloaded because audio *attenuation occurs after the* product detector. Clearly, some form of i-f gain control is the answer; and for weak-signal cw reception, this may be in the form of a bias control in the final i-f stage.

Also not to be overlooked is the fact that a product detector is a mixer (by definition a nonlinear device), and the less noise or excess signal presented to it the better. The gain of the incoming signal, therefore, should be reduced before the signal reaches the product detector.

A receiver designed for truly weak-signal reception must reduce the strongest signal to near zero amplitude while providing sufficient amplification to obtain usable audio response from a weak signal.

tests

I ran some tests to determine ways of improving signal-to-noise performance. A block diagram of the receiver configuration for these tests is shown in fig. 1. Note that the rf stage has neither manual nor automatic gain control but has regeneration control available. The buffer and 465-kHz i-f have agc plus an overriding manual control; the main control operates on the cathodes of the two 85-kHz i-f amplifiers, which have no agc. Two audio filters are also available.

Five controls are shown. Details of each are given together with information concerning the 85-kHz channel and the audio-output meter.

control 1. The rf stage is regenerative to the point of oscillation at maximum setting. At minimum setting regeneration ceases, and the amplifier operates at normal gain.

control 2. This control replaces the usual rf gain control. It provides cathode-bias control for two stages, both of which may be controlled by agc. Control 2 normally runs wide open but is adjusted to attenuate strong signals. (An rf stage and its mixer are seldom overloaded; the stages following the first mixer are those susceptible to overload.)

control 3. This control can take the Q multiplier *into oscillation.* The *Q multiplier* is used as a peaking device rather than a selectivity control. The 1.5-kHz bandwidth of the 85-kHz i-f channel provides the i-f selectivity.

control 4. Cathode bias for the two 85-kHz i-f stages is controlled by this pot, which replaces the usual audio gain control.

control 5. This is a 2-pole 3-position

switch that enables either one or both audio filters. The filters are toroid LC tuned circuits, one in the product-detector output and the other in the grid circuit of the first audio stage. Resonant frequency is 1 kHz, and the filters provide considerable gain. A suggested schematic is shown in **fig. 1**.

audio-output meter. The audio-output meter is designed for a 600-ohm receiver output; zero dB represents 1 mW. The lowest-scale reading is -20 dB (10 microwatts). The point of zero signal with sensitive headphones is about -7 dB. It is necessary to calculate responses below -20 dB, as we shall see.

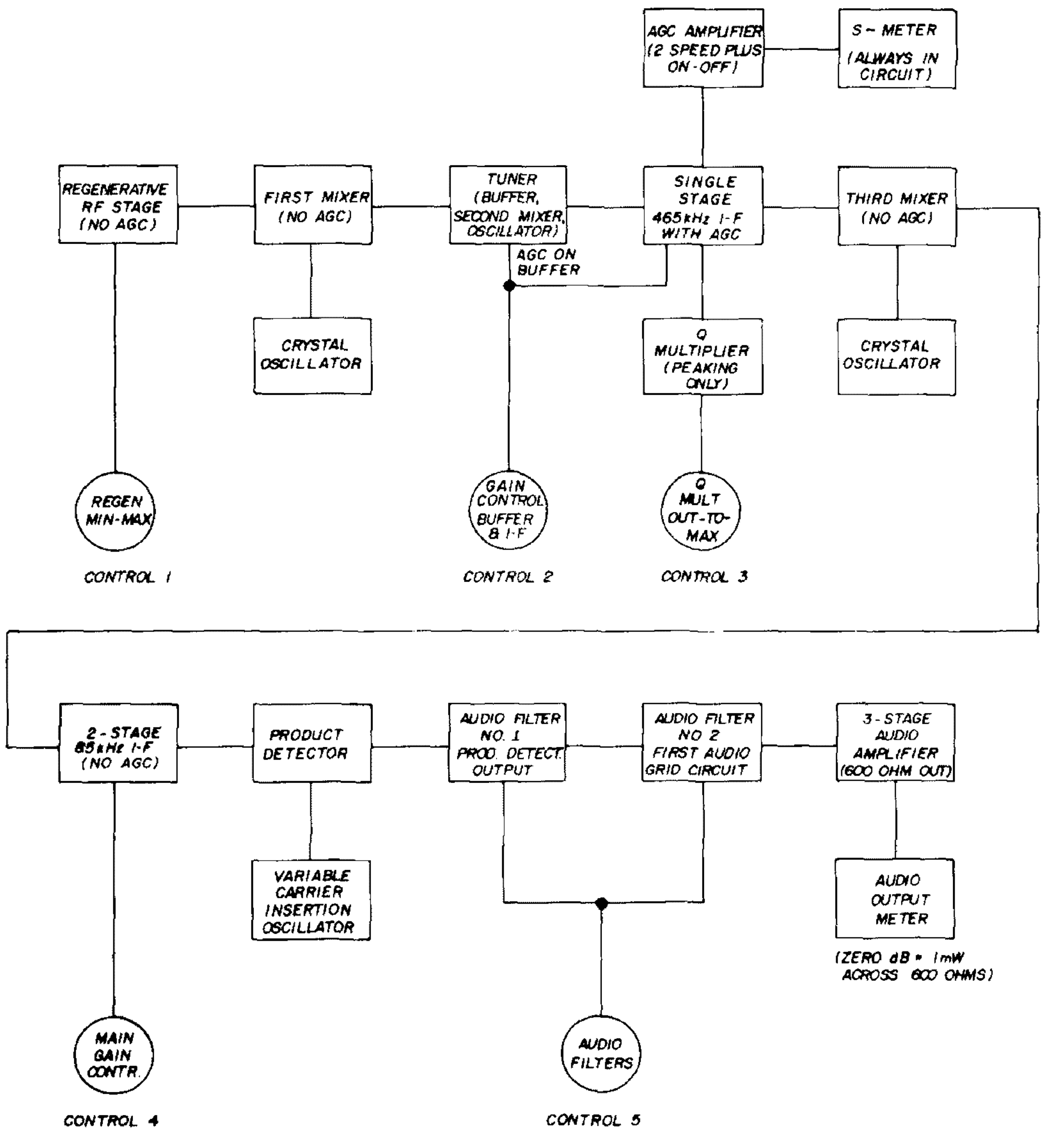


fig. 1. Low-noise receiver that produced the test results in table 1. Control functions are described in the text.

85-kHz i-f channel. The input to this channel is two cascaded i-f transformers. A total of four undercoupled transformers give a bandwidth of about 1.5 kHz.* The carrier-insertion oscillator is variable from 83 to 87 kHz and is adjusted to the audio-filter requirement.

*The 85-kHz i-f strip can be a BC-453 war-surplus aircraft receiver (tuning range 190-550 kHz). The shape factor of this strip is quite good: 1.5 kHz at 6 dB down and 6.5 kHz at 60 dB down. It is still available; try G & G Radio Electronics Co., 45-47 Warren St., (2nd Floor), New York, N. Y. 10007. Price used is $16.95 complete with tubes. The set can be easily transistorized. editor.

preparation for tests

Instead of a signal generator I used a live signal via the antenna. The signal source was a shielded low-power oscillator located clear of receiver and antenna. The output was adjusted for a signal 10 dB over the *basic* internal noise level. The signal showed no level variation at either the S meter or audio output.

The antenna was arranged so I could switch from receive to dummy load to compare signal level with noise at each setting. Noise via the antenna was not taken into account, but on the preliminary test I noted that, in the absence of a signal, the noise increased by less than 5 dB when changing from dummy load to receive.

basic adjustments

Agc was switched off, and control 2 was set to maximum. Regeneration was at Minimum, the Q multiplier was OFF, and both audio filters were OUT. The output meter was placed on the scale -20 to +17 dB, and control 4 was adjusted (dummy-load position) to give a noise output of -10 dB. When switching to receive and tuned to the test signal, the output increased to zero dB; in fact, the test signal was adjusted to produce this level.

At this point, the receiver compared well with an average communications receiver. A check over the cw band showed a lively response. The test-signal level, in microvolts, is not known. It is irrelevant to the tests, but it certainly did not exceed 1 microvolt.

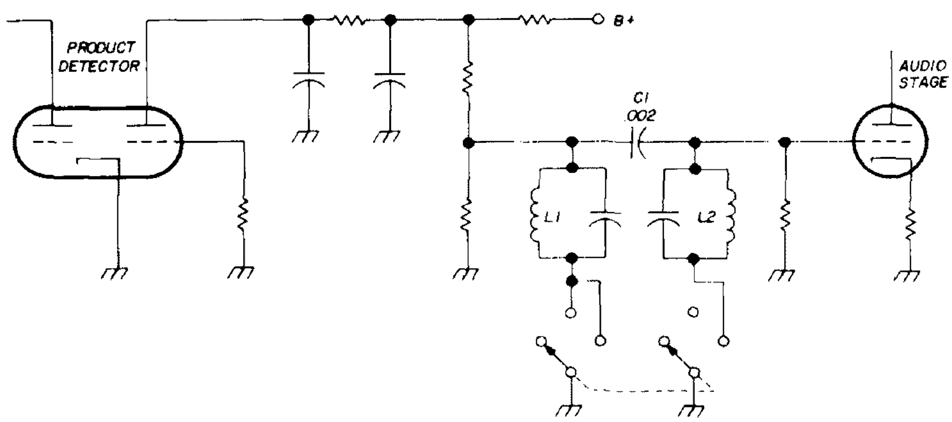


fig. 2. Suggested audio filter. L1, L2 may be 88 mH toroids or preferably 300 mH, which have a better response curve and higher gain. C1 should be as shown if it is now larger in your receiver.

test objectives

It was desired to start with basic adjustments approximating those of an average good receiver. A fade-free signal would be available to produce exactly 10 dB above receiver internal noise. A record would be kept of the following effects on receiver operation:

1. Audio filtering.
2. Regeneration.
3. Q multiplier.
4. Combinations of the above.

test results

Results of the tests are shown in **table 1**. I was surprised by the data even though I'd had considerable experience with the receiver and knew its capabilities. A study of **table 1** leads to the following conclusions:

audio filtering. A 20-dB improvement for one audio filter proves that audio filtering really reduces noise. The fact that the second filter gave only 5 dB additional improvement indicates that the filter skirt-width doesn't greatly improve after the first big bite at the apple.

regeneration. The 8-dB improvement shows that the signal does increase faster

than the noise when regeneration is used. The increase in sensitivity is worthwhile.

Q multiplier. This is really a form of highly stable i-f regeneration. Selectivity measurements have shown that passband sharpening is completely nullified if the band being sharpened is wider than a subsequent channel. The 465-kHz i-f channel, even with the Q multiplier in use, is wider than that of the 85-kHz

analysis of test results

The tests show that improved sensitivity reduces noise because less subsequent amplification is required – in other words, the gain may be reduced. Also, restricting audio bandwidth is of great assistance. Moreover, it is clear that a multi-tube receiver can be made virtually noise-free. The actual amount of noise may be estimated from the following discussion.

table 1. Test results using a "live" signal from the antenna. Signal source was an oscillator adjusted to produce a signal 10 dB over the basic internal noise level of the receiver. See fig. 1 and text for control position.

rf regeneration	out	out	in	in	out	out	in	in
Q multiplier	out	in	out	in	out	out	in	in
first audio filter	out	out	out	out	in	in	in	in
second audio filter	out	out	out	out	out	in	out	in
noise level (dB)	-10	-10	-10	-20	-20	-20	-20	-20
test signal level above noise (dB)	10	13	18	30	30	35	37	40
improvement (dB)		3	8	20	20	25	27	30

channel. Nevertheless, the 3-dB improvement is most effective, as shown in the next paragraph.

regeneration + Q multiplier. An improvement of 20 dB is identical to that obtained with a single audio filter. The difference between the two conditions is merely one of bandwidth. However, it is desirable to retain this improvement because it's not always necessary to use audio filters. The receiver is invariably used with both regeneration and Q multiplier at maximum.

regeneration + Q multiplier + filters. With the noise reduction already obtained, it is obvious that audio filters would have less work available. In fact, the single filter now contributes only 7 instead of 20 dB, and the second filter adds a mere 3 dB improvement. At the same time, the *total* improvement is a very worthwhile 30 dB.

For headphone reception an audio output of 100 microwatts is more than adequate; 1 to 2 milliwatts will drive a speaker for use in a small room. If we convert the dB numbers into audio power, we obtain a better perspective (table 2).

Working from table 2 and referring to table 1, we can back off the gain control (in theory) until the test signal is 100 microwatts (-10 dB) and estimate the amount of noise that will be present. At the *basic* setting, with the signal at -10 dB instead of zero dB, the noise level will still be 10 microwatts – a considerable amount.

For a 20-dB improvement, the noise will decrease to 0.1 microwatt, at which level it barely intrudes on the ear. For a 30-dB improvement, the noise will be a mere 0.01 microwatt. However, if we back off the gain until the signal is 10 microwatts (-20 dB), the noise will be

0.01 microwatts for a 20-dB improvement and 0.001 microwatts for a 30-dB improvement.

Until now, we've considered a signal which was, basically, only 10 dB above the noise – really an S1 signal. Let's now consider a signal 20 dB above the noise, improved 30 dB and reduced to produce 10 microwatts. The noise is now -70 dB or 0.0001 microwatt.

Now we'll add one point for good measure: a 0.1-microwatt signal from a loudspeaker is barely audible at 6 feet, but a 10-microwatt signal can be *copied* at 30 feet or more.

table 2. Data for estimating improvement in receiver response to noise based on the combinations in table 1.

audio meter (dB)	output power (μW)	remarks
0	1000	very strong on headphones
-10	100	strong headphone signal
-20	10	comfortable headphone signal
-30	1	weak headphone signal
-40	0.1	a tolerable noise level
-50	0.01	an excellent noise level
-60	0.001	noise barely noticeable
-70	0.0001	virtually zero noise

conclusions

Admittedly the tests involved only signal versus internal noise and the results will be degraded by external noise. However, the use of audio filters narrows the audio spectrum so that external noise is greatly reduced.

The fact that a triple-conversion receiver was used is incidental. Communications receivers have an rf stage, and many have a Q Multiplier. Audio filters may be used with any receiver that has an audio stage. Triple conversion doesn't offer much advantage over a double-conversion receiver if sharp i-f filtering is available, but triple conversion does permit rejection of the audio image without filters (giving single-signal response); and triple conversion permits agc voltage generation and the use of an S meter at the second i-f while bfo/cio injection and gain control follow the third i-f stage.

The amount of audio amplification provided doesn't appear to affect noise too greatly because we seldom use all the audio amplification, although we use much more on weak signals. Nevertheless, we should remember that the audio amplifier has little or no internal noise – in the sense that noise from audio tubes is not amplified to the same extent as that from earlier tubes, and there is less loading of the product detector where adequate audio amplification is available.

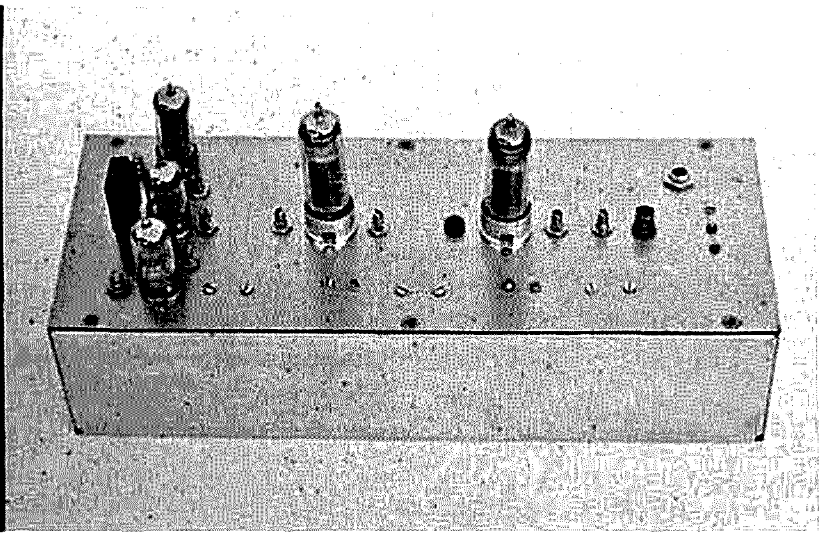
Although 1 to 2 milliwatts of audio usually suffices for cw operation, it is advisable to consider the potential audio outputs available from a sensitive low-noise receiver. Based upon an output of 1 milliwatt for zero-dB, +10 dB is 10 milliwatts and +30 dB is 1 watt, with +40 dB increasing to 10 watts!

Where noise level stands at -20 dB, a signal 20 dB above noise is a mere 1 milliwatt, but at 40 dB above noise we have 100 milliwatts. If this is the type of output we expect from a 1-microvolt input (see table 1) what can we expect from a local signal some 60 dB stronger? This would be 100 dB above the noise; a potential output of one *kilowatt* of audio for -40 dB of noise! To reduce such a signal to 1 milliwatt calls for the noise to go back to -100 dB; i.e., to one ten-millionth of a microwatt. This is the measure of gain reduction we need if our receiver is *really* sensitive with low-noise response.

Even a regenerative rf stage will handle the 60-dB increase (from say 0.1 microvolt to a volt or two); but the agc and/or rf gain must reduce early amplification considerably; and finally we need considerable attenuation later in the receiver. We can, however, visualize the amount of signal passing through the product detector if we rely on an audio-input attenuator running almost closed off to limit the audio to 1 milliwatt or less, even after reducing the rf gain.

There is nothing more pleasing to the cw DX man than a really weak signal producing a 1-milliwatt audio signal with an absolutely quiet background.

ham radio



low-cost 220-MHz exciter

This simple crystal-controlled exciter for 220-MHz provides several watts CW output — the final stage may be modulated if desired

Martin Beck, WB6DJV, 1637 Hood, Wichita, Kansas 67203

When an amateur contemplates 220 MHz the first thing that enters his mind goes something like this, "If I build a decent exciter for 220 a lot of hard work and time will very likely go down the drain; I have a pal with a surplus RDZ and he says there's nothing going on up there anyway." Well, there *is* activity on 220. And the RDZ doesn't have a low-noise 2N4416 front end. As to the matter of the exciter troubles — here is a little gem that uses a minimum of components, does not have to be neutralized, keys like a dream and uses five garden-variety tubes.

The exciter consists of a 6AQ5 oscillator stage, followed by three triplers and a 220-MHz power amplifier (fig. 1). I chose 6C4s for the first two triplers because they are low cost, and RCA rates them as 5-watt oscillators at 144 MHz; they are extremely efficient at 24 and 72 MHz.

construction

When you wire the tripler sections you will need a small soldering pencil to work in the cramped quarters (I used a Weller WPS). Don't be discouraged by the compactness of the first three stages. Lay in the heater and B+ wiring first, then install the bypass capacitors. Next install all resistors and capacitors. Don't cheat by using wire; you will only need the component leads.

When you have the first three stages wired, plug in the 6AQ5, a suitable 8-MHz crystal,* apply power, set the oscillator tuning capacitor half open and screw in the L1 coil slug until you hear the oscillator signal on a general-coverage receiver. Now plug in the first 6C4 tripler, set the plate-tuning capacitor half open, and adjust the L2 tuning slug for maximum 24-MHz signal indication. Plug in the second 6C4 tripler and go through the same procedure, tuning for maximum signal at 73 MHz. These initial settings will put you in the ballpark when you are ready to line up the last two stages.

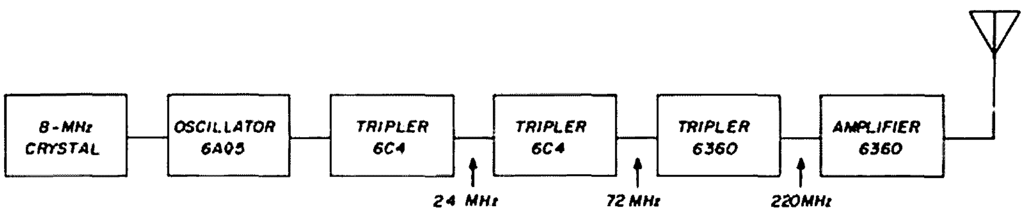
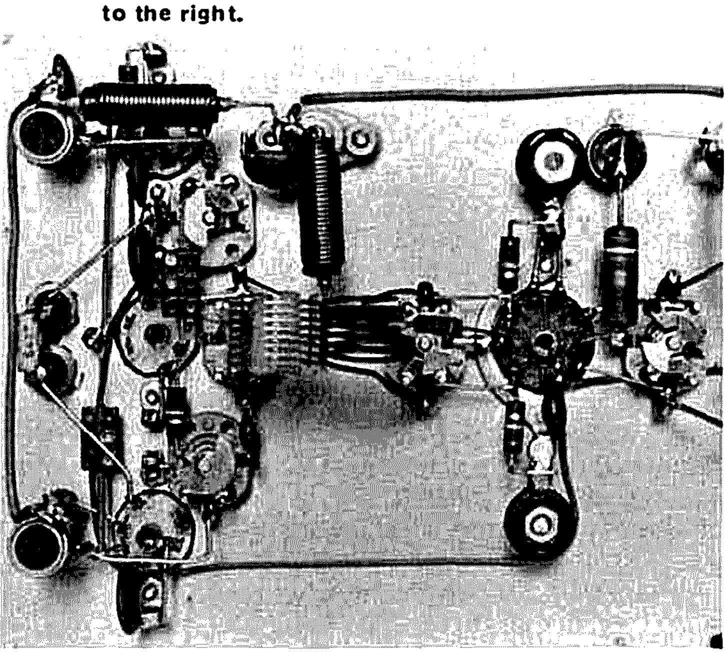


fig. 1. Block diagram of the simple 220-MHz exciter.

Bottom view of the exciter showing the crystal oscillator and 6C4 tripler stages to the left; 73- to 22-MHz 6360 tripler is to the right.



photos by M. McConnell

Turn off the power, remove the three tubes, and wire the third tripler and final power amplifier stages. After checking your wiring for errors mount the coils on the variable butterfly capacitors. To tune up these stages the coils are set on frequency by either pinching them shut or spreading them apart (with the tuning capacitors half open). If you use 222.5 as your tuneup frequency the variable capacitors will cover the entire 220-MHz band.

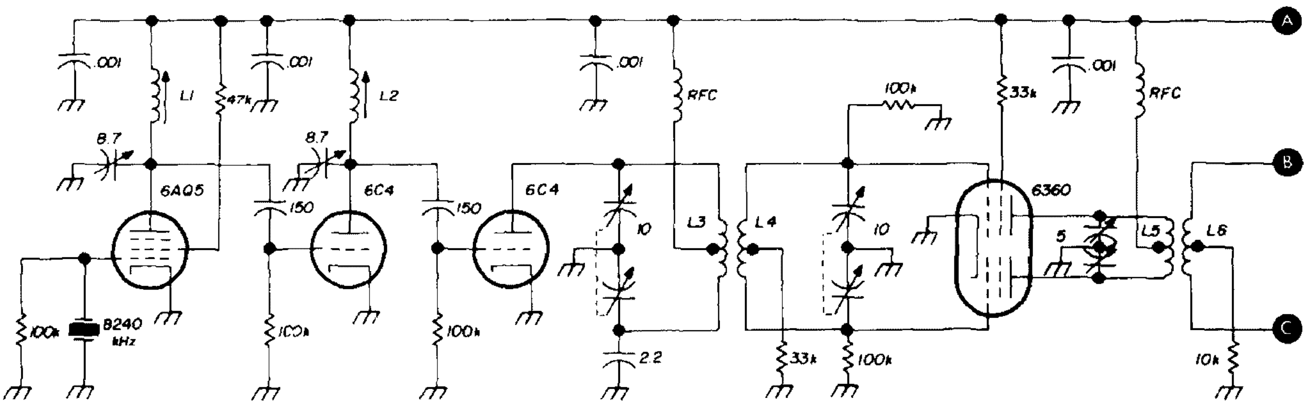
Do not change the value of the 6360 screen resistors. With the values shown in fig. 2 you will not have to measure plate current as power input is within safe limits when the circuit is tuned to resonance. In rigs of this type I have always preferred to read the grid current of each stage except the final power amplifier. If you want to put in meters and meter switches, fine, but it's certainly not a necessity.

final tuneup

When tuning up the 6360 power amplifier, open the B+ line to *both* the screen and plate. When you have obtained maximum possible final grid drive, turn

*For output on 222.5 MHz, select a 8.240 MHz crystal; the first tripler will be at 24.720 MHz and second tripler, 74.160 MHz. For 220 MHz use a 8.148-MHz crystal; a 8.333-MHz crystal will put output at 225 MHz.

the power off and reconnect the plate and screen voltage. Apply power again, tune the plate circuit for maximum 220-MHz output, using both the butterfly color whatsoever. If *one* of the tube plates begins to show color turn off the power and check your tank-circuit wiring for symmetry. Make sure the mechanical



C8,C10,C13	10.8-pF butterfly (E. F. Johnson 160-211)	L5	5 turns no. 16, 3/8" diameter, 1¼" long
C11	5.1-pF butterfly (E. F. Johnson 160-205)	L6	5 turns no. 16, 3/8" diameter, 1-3/4" long
L1	16-29 μH (J. W. Miller 4407)	L7	5 turns no. 16, 5/16" diameter 1" long
L2	3 μH (J. W. Miller 4608)	L8	1½ turns no. 18 enameled, closewound, 7/8" diameter
L3	8 turns no. 18 enameled, 5/8" diameter, 1/2" long		
L4	7 turns no 18 enameled, 1/2" diameter, 3/8" long		

fig. 2. Low-cost 220-MHz exciter uses 8-MHz crystal oscillator, three frequency triplers and power output stage. No neutralization is required.

and series-link capacitors.

Now, turn off the room lights and carefully observe the plates of both 6360 tubes. There should be absolutely no layout of the 6360 stages is symmetrical; electrical symmetry will follow. If you find you must neutralize the 6360 final, you did something wrong.

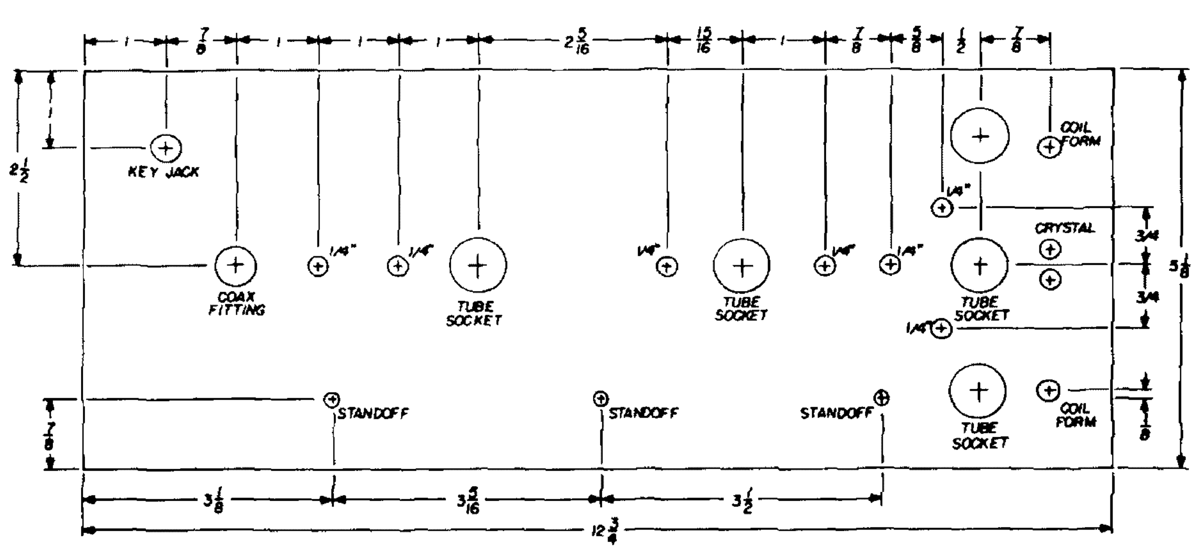
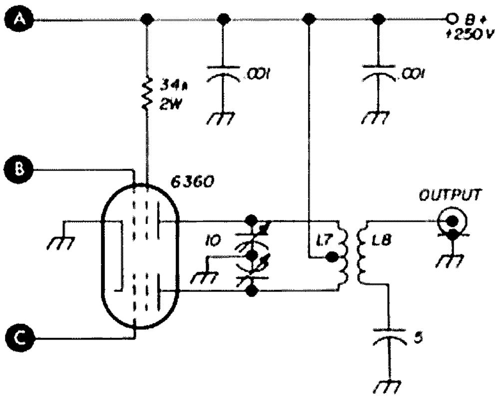


fig. 3. Layout for the 220-MHz exciter.

The bias on the 6C4 and 6360 control grids is provided by drive from the previous stage. If drive fails every tube in



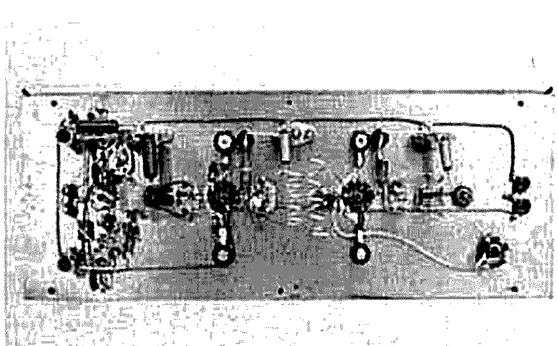
the line will be quickly destroyed. This has never happened to me but I'm always on the lookout for drive failure. You can put a fuse in the B+ line, but if you tune slightly off-resonance you'll lose the fuse – which can be annoying, and always happens at the worst possible time. Obviously, a separate bias supply is the best approach, but this little rig was designed for utter simplicity and low cost, so none was included.

If you're a CW man as I am you're all set with a few watts out on 220 MHz. If you want more power you can build the 220-MHz power amplifier described in an earlier issue of *ham radio.*[1]

reference

1. Martin Beck, WB6DJV, "An RF Power Amplifier for 220 MHz," *ham radio,* January, 1971, page 44.

ham radio



Bottom view shows layout of the exciter.

the simplest
cw audio filter

This most simple of all cw audio filters series tunes the built-in inductance of your headphones

Gene Hubbell, W7DI, 6633 East Palo Verde Lane, Scottsdale, Airzona 85251

For the simplest possible cw audio filter all you have to do is put a capacitor in series with your high-impedance headphones. Lower impedance headphones may require additional inductance, but in most cases a series capacitor will resonate the headphones to a convenient audio frequency.

I have used this effect for some time, and recently set up the test circuit in fig. 1 to determine the proper capacitance values for various types of headphones. I used an old Western Electric audio signal generator with 600-ohm output into a 500:3.2-ohm audio transformer. A sensitive vtvm across the headphones measures circuit voltage. For low-impedance headphones a surplus 88-mH toroid inductor may be switched into the circuit.

With the headphones connected directly to the 3.2-ohm winding of the transformer the voltage was approximately 0.22-volt rms. With a series capacitor, and the audio generator tuned to the resonant frequency, the voltage across the headphones increased to as much as 0.7 volt; this represents considerable audio gain.

At frequencies below resonance response falls off rapidly due to increasing capacitive reactance. Above resonance the voltage drops slowly to the 0.22-volt level. However, when the 88-mH toroid is in the circuit response drops much more sharply at frequencies above resonance.

High-impedance headphones work best in this simple circuit. Medium-impedance headphones do fairly well, and low-impedance headphones (6 to 8 ohms) exhibit little, if any, series-resonance. You can get an idea of the impedance range of your headphones by measuring their dc resistance. If the dc resistance is 1000 ohms or more they can be classified

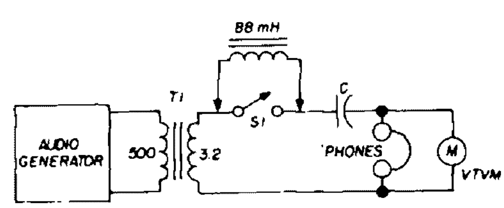


fig. 1. Test circuit for checking resonance of series-tuned headphones.

table 1. Performance of various headphones with series-resonant capacitors.

headphones	peak frequency (Hz)	capacitor (μF)	peak voltage
Western	580	.015	0.62 V
Electric	750	.01	0.62 V
type 194W	800	.0075	0.60 V
(2000 ohms dc)	1150	.005	0.55 V
Murdock Signal	600	.05	0.55 V
Corps type R-14	700	.02	0.50 V
(1700 ohms dc)	1000	.015	0.65 V
	1200	.01	0.68 V
	1350	.0075	0.66 V
Telex	750*	.22	0.37 V
(225 ohms dc)	800	.32	0.40 V
	900	.15	0.40 V
	950	.22	0.40 V
	1050*	.1	0.40 V
	1150	.15	0.40 V
	1300	.1	0.37 V
Murdock type	750*	.22	0.35 V
P-23 CAATC	900*	.22	0.37 V
(130 ohms dc)	1050*	.15	0.45 V
	1300*	.1	0.45 V

*88-mH toroid inductor in circuit. Note dual resonance of Murdock P-23 headphones with 0.22-μF series capacitor.

as high-impedance types; 150 to 300 indicates medium 500- to 600-ohms impedance. The dc resistance of low-impedance headphones will usually be much less than 100 ohms.

Some of the results I obtained are shown in **table 1**. The Western Electric 194W (antiques) and Murdock Signal Corps R-14 headphones are high-impedance types: the Telex and Murdock P-23 headphones are medium impedance. The resonance effect of the 88-mH inductor is evident in the two tests made with the Telex headphones.

Since most of the headphones used by amateurs were not originally designed for flat response one or more peaks usually show up as you change the audio signal frequency. If the series capacitance is chosen carefully the tuned peak can be made to coincide with a natural signal peak, thereby accentuating the cw filtering effect.

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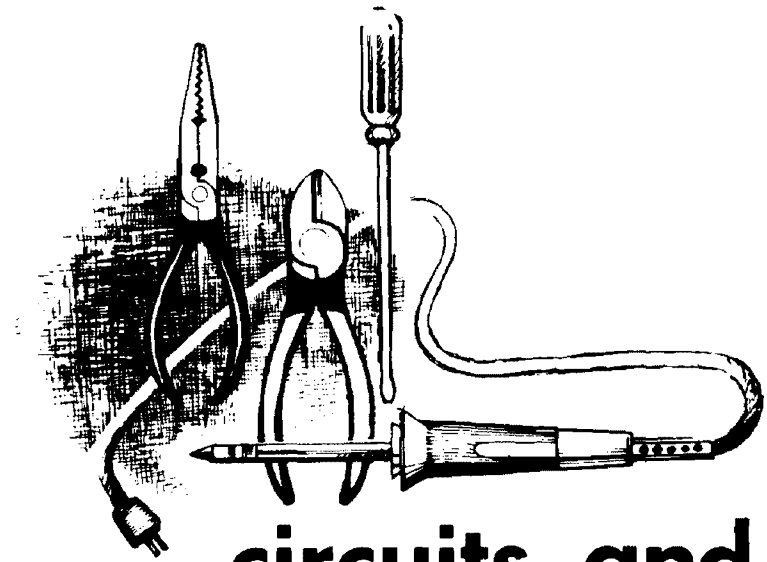


circuits and techniques

ed noll, W3FQJ

direct conversion receivers

Solid-state direct conversion is encouraging receiver construction; simplicity, low cost and good performance are its virtues. QRP fans are properly enthused because these receivers match the size, cost and battery-drain limitations of their low-power operations. It is not unusual to work other hams operating with substantial transmitter power but receiving on a small direct-conversion receiver.

Watch out for considerable commercial activity too; this mode of reception is a specialty of the Ten-Tec Company. Direct-conversion receivers match and often out-perform many of the commercial receivers in the $75 and $175 class, especially if these are single-conversion types.

theory of operation

In a direct-conversion receiver the signal is applied directly to the detector along with a locally generated carrier of similar frequency. The output circuit emphasizes the difference-frequency components, these being the modulation on the incoming signal. A direct conversion is made from signal frequency to modulating frequency. A more exact definition for a direct-conversion receiver would be a receiver that converts directly to the modulating frequency by heterodyning.

The demodulation is handled by a product mixer. In the mixing and multi-

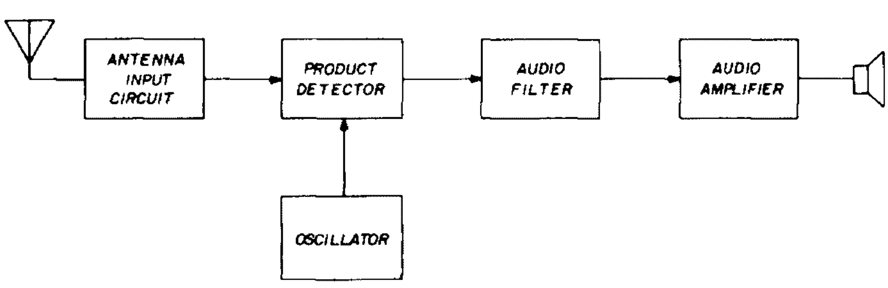


fig. 1. Essential details of a direct-conversion receiver.

plying of signal and oscillator frequencies many frequencies are produced in the output. The one of concern is usually the difference frequency. When the signal and oscillator frequencies are the same, a special form of product mixer, the product detector, is used.

Leo Gunther, VK7RG, in the April 1970 issue of the *Australian EEB*, states it this way, "The oscillator frequency in the product mixer need not be the same as that of the signal; though it is when used as a product detector."[1]

Inasmuch as the product detector form of linear mixer produces a family of output frequencies, the output circuit must include an audio filter that filters out undesired components and emphasizes the original modulating frequencies. Finally, the power level of the demodulated audio must be increased to drive a headset or loudspeaker.

The gain of the receiver is determined by the gain of the audio amplifier. If good signal-to-noise ratio and sensitivity are to be obtained, a low-noise high-gain audio amplifier is essential.

A single-sideband signal is demodulated by tuning the oscillator frequency to the absent carrier frequency of the incoming signal. A CW signal is demodulated by off-setting the receive oscillator frequency from the incoming carrier frequency. A conventional a-m signal is demodulated by tuning the oscillator frequency to the incoming carrier frequency using the same zero-beat technique used when demodulating an a-m signal with a sideband receiver.

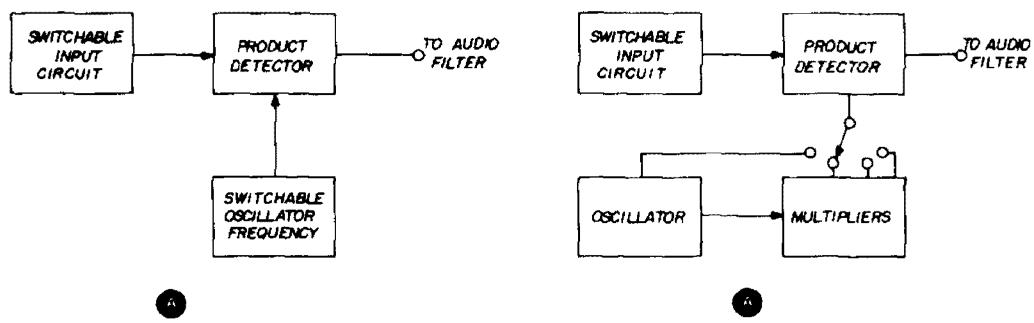


fig. 2. Two schemes for multiband operation of direct-conversion receivers.

what constitutes good direct-conversion practice?

The conversion should be linear. A balanced detector circuit with good linear performance over sufficient dynamic range is preferred. Hot-carrier diodes with their low noise content and low intermodulation distortion are popular. The field-effect transistor (both dual-gate and dual-fet types) is attractive for the same reasons. However, as VK7RG points out, the low signal levels at which the normal product detector operates in a direct-conversion receiver is an important factor in the low cross modulation performance.

The usual direct-conversion receiver includes no rf stage. None is required when the product detector itself has a very low noise level. Selectivity is determined by the output audio filter. Any rf stage ahead of the product detector has a marginal influence on the overall selectivity. An overpowering local station may give some trouble, although its effect is likely to be less than in the usual solid-

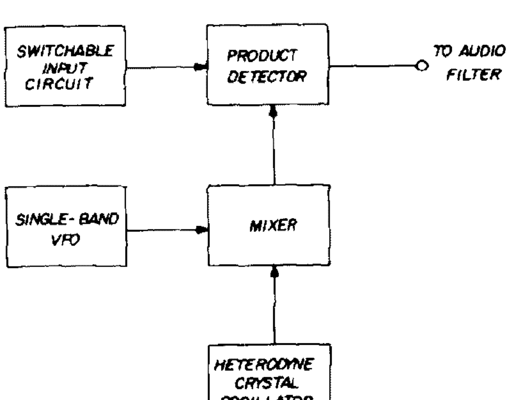


fig. 3. Using heterodyning to obtain multiband operation with a direct-conversion receiver.

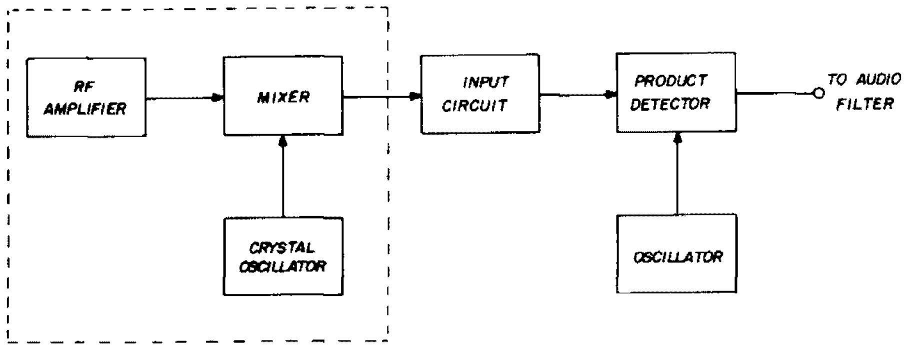


fig. 4. An outboard tuner can be used with a direct-conversion receiver to provide multiband operation.

state superheterodyne receiver. A narrow-band filter is appropriate for CW; wider band (300-2500 Hz), for ssb reception.

Presently the audio filters used in direct-conversion receivers are relatively simple types. The inclusion of audio filters with better shape factors is just a matter of time.

Sensitivity is influenced by the overall gain of the audio amplifier and the noise content of the detector and audio input stage. A low noise fet input stage seems expedient not only because of its low noise content but also because its high impedance permits the design of a more simple good-shape audio filter.

The oscillator must be very stable. However, this requirement is now old-hat because the same order of stability is needed in more conventional sideband receivers and in transmit or transceiver vfos.

There are three ways of adding bands to a basic single-band direct-conversion receiver. The obvious procedure is shown in **fig. 2.** In example A, a multiband oscillator is used. You need only to switch oscillator frequency and antenna input resonance to change bands. An alternative to this scheme is to use a stable low-frequency oscillator and a switchable multiplier chain that generates the harmonics needed for operation on higher-frequency bands.

An idea used by C. F. Dorey[2] employs a stable vfo and one or more crystal oscillators, fig. 3. A mixer selects either the sum of difference frequency to obtain an appropriate injection frequency for the desired receive band.

The final method,[3] shown in **fig. 4,** changes the direct-conversion receiver over to a superhet for reception on other bands. In this case individual band converters are placed ahead of the product detector. These outboard tuners can consist of a crystal oscillator-mixer combination or an rf amplifier and mixer-oscillator.

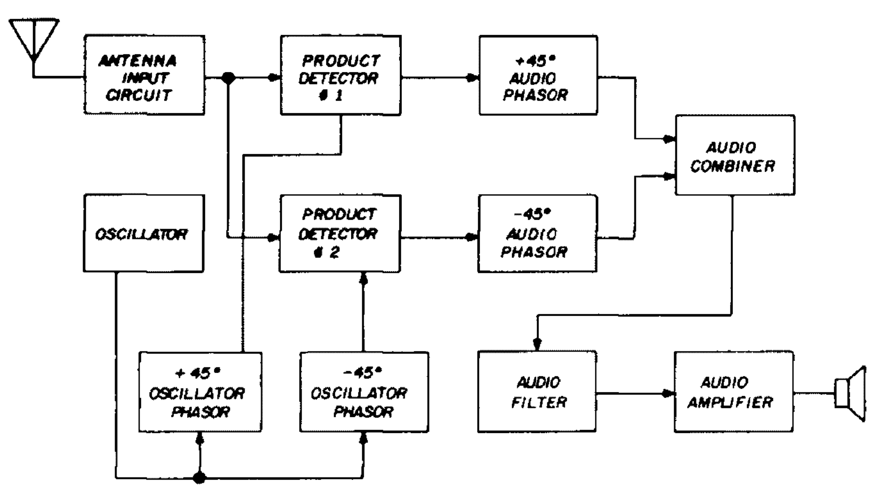


fig. 5. Two-phase direct-conversion receiver.

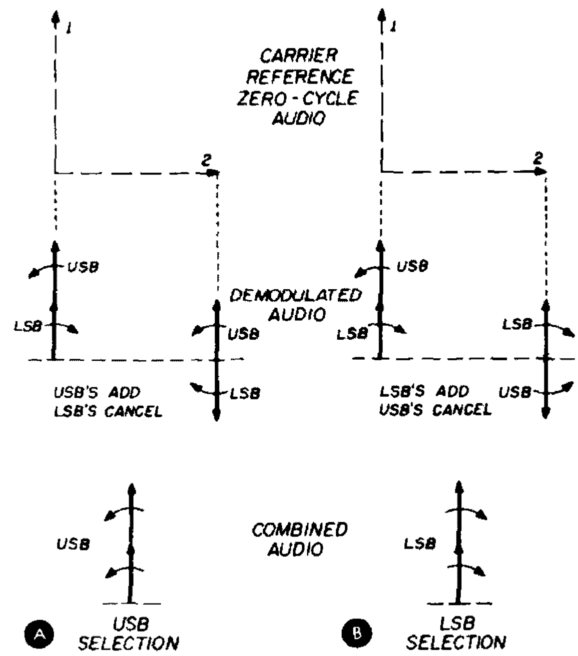


fig. 6. Vector diagrams of two-phase direct-conversion receiver signals.

adjacent channel interference

The conventional sideband receiver, be it used for CW or sideband reception, includes a sideband filter that rejects carrier and undesired sideband spectra. The usual direct-conversion receiver has no such facility. If you are tuning in a lower sideband signal, an adjacent channel signal may beat with the oscillator frequency to produce audio difference components that fall in the upper sideband spectrum. These will be heard in the output. Radio-frequency selectivity is of little use in rejecting these opposite-side audio images. Good audio filter selectivity is of some help but this can be carried to the point at which the desired modulation loses intelligibility and/or imposes an impractical pinch on oscillator stability.

two-phase direct conversion

A double-balanced product detector[4,5] with appropriate phasing can be used to cancel out image-side components. This adds complexity as shown in **fig. 5.** However, integrated circuits are good devices for designing double-balanced systems with little increase in weight and size. The added stages are worth the effort if you strive for performance that will match more conventional high-quality sideband receivers.

In the two-phase system the incoming signal is applied in phase to both product detectors. Oscillator components are applied 90° out of phase. The two demodulated audio signals are then made 90° related with an audio phase-shift network. Next the two signals are combined prior to application to the main audio filter and a follow-up audio amplifier. What happens is shown in the vector diagrams of **fig. 6.** When adjusted for upper sideband reception the audio frequency components demodulated from one side of carrier frequency are additive. Conversely, any audio frequency components demodulated from the other frequency side of carrier are subtractive.

Incorporated switching selects either upper or lower sideband reception. Switching can take place in the combiner by a change-over between difference or sum operation. Two-phase operation may be the next general forward step in direct-conversion techniques.

cw transceive

Direct conversion is just fine for transceive operation. One oscillator (crystal,

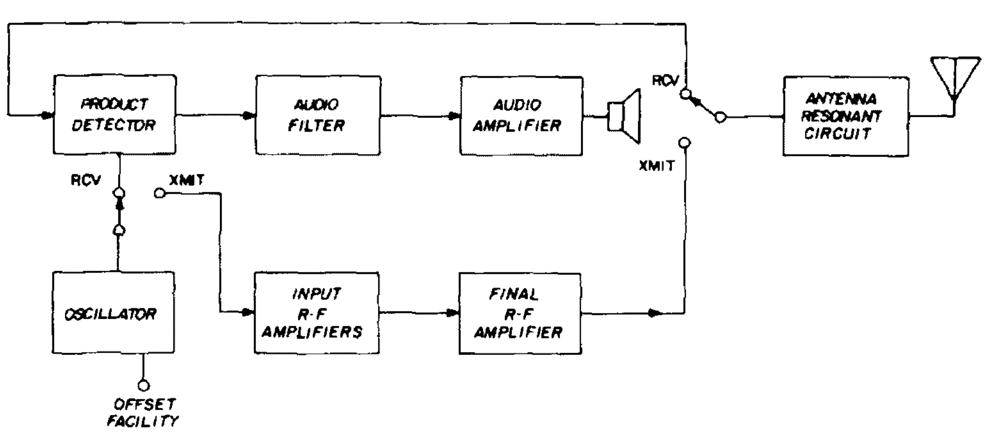


fig. 7. Transceiving with direct conversion.

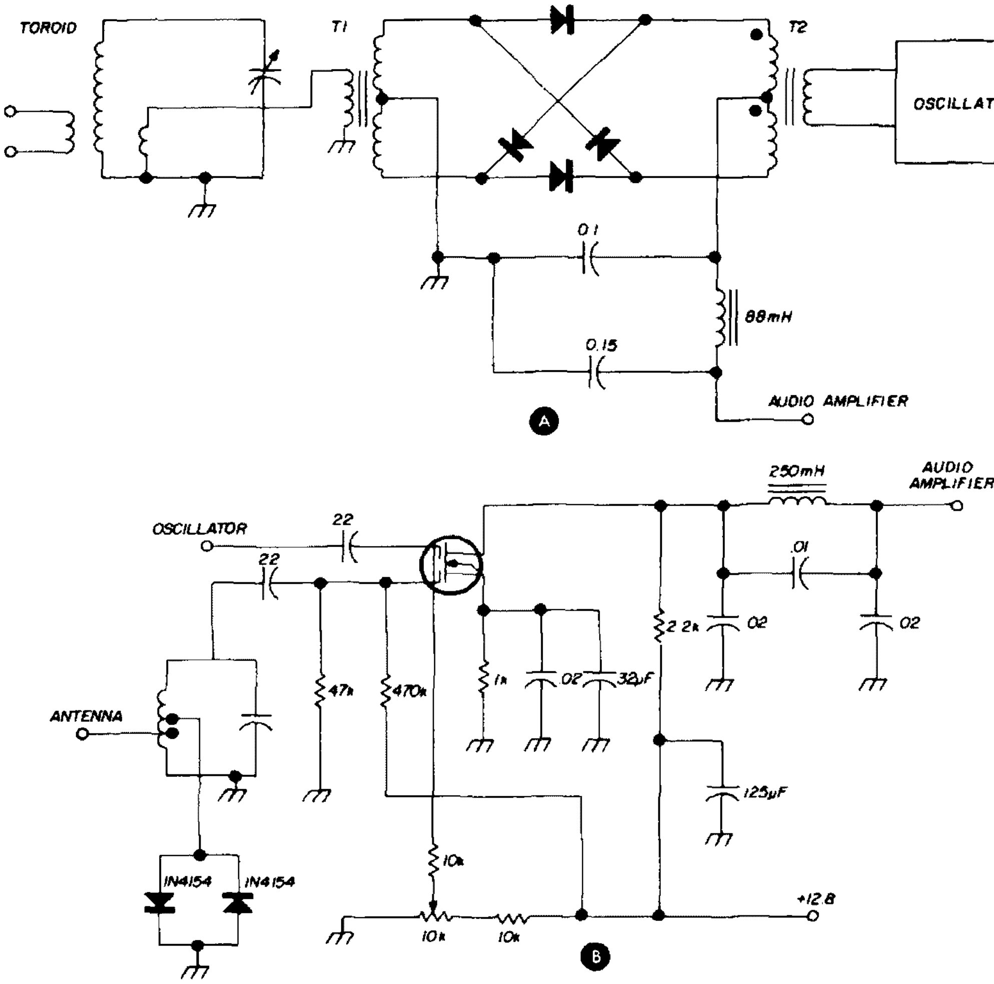


fig. 8. Two popular product detectors. Hayward-Bingham circuit is in (A). Transformers T1 and T2 are trifilar-wound toroids. Circuit in (B) is used in the Ten-Tec RX10 direct-conversion receiver.

vxo or vfo) serves both receiver and transmit modes, fig. 7. There is a slight difference between the transmit oscillator frequency and the desired offset frequency operation of the oscillator for CW reception. A correction can be a part of the transmit-receive switching[6, 7, 8, 12] in the oscillator circuits. You can then adjust the tone of the demodulated CW without changing the frequency of the

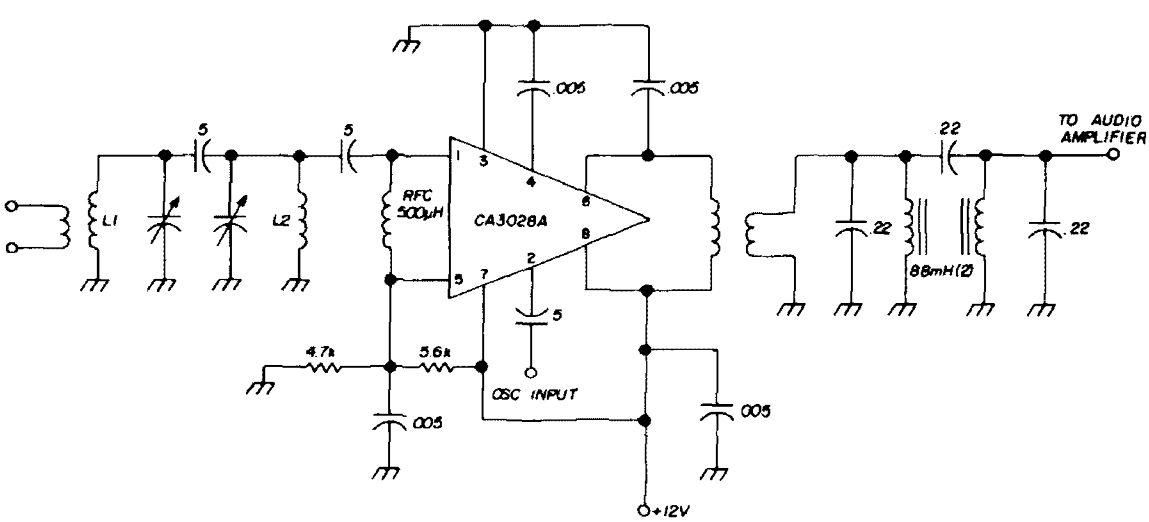


fig. 9. Integrated-circuit product detector uses RCA CA3028A. L1 and L2 are mutual-C coupled toroids.

transmit oscillator. The same preset switching arrangement also compensates for any change in oscillator frequency that may result from loading shifts in switching between transmit and receive.

typical circuits

The product detector is the key stage of a direct-conversion receiver because of its influence on distortion, cross modulation, noise and sensitivity. Low-noise devices and a balanced circuit keep noise and cross modulation at a low level. If the product detector is somewhat less than the best it can be preceded by an fet amplifier that will improve the signal-to-

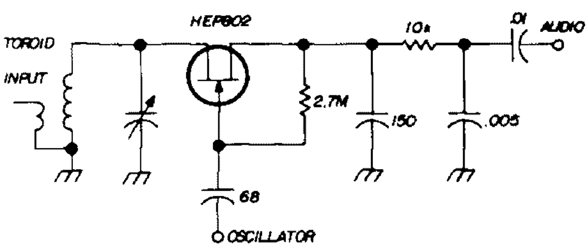
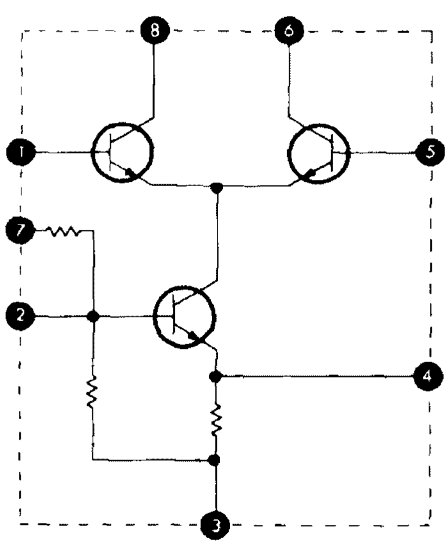


fig. 11. Simple product detector and audio filter used in receiver designed by K1BQT.

noise ratio and, to a limited degree, the selectivity as well.

The two most popular forms of product detector are shown in **fig. 8**. Example A is the common balanced-ring arrangement using the low-noise hot-carrier diodes.[9] This circuit customarily uses two broadband trifilar input and output windings. Link coupling is used between the input of product detector and the antenna resonant circuit.

fig. 10. Circuit diagram of the CA3028A differential-amplifier ic.



Other types of low-noise diodes are used and perform well. For single-band

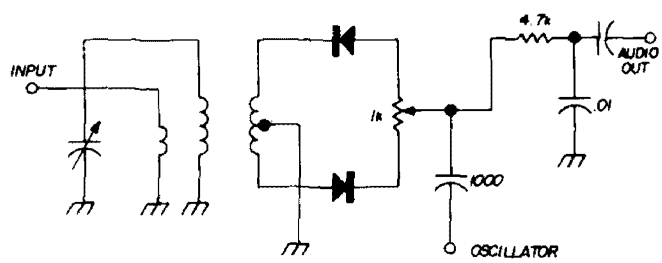


fig. 12. Simple two-diode product detector devised by PAØKSB.

operation a resonant input circuit can develop a higher output and improve the signal-to-noise ratio.

The dual-gate mosfet provides a fine low-noise and minimum cross modulation product detector. The circuit of **fig. 8B** is used in the popular Ten-Tec RX10 direct-conversion receiver. One gate is for signal injection; the second gate, for oscillator injection. A switchable resonant input circuit permits operation on four bands, 15 through 80 meters. A 2-kHz audio filter is located between the drain and the input to the audio amplifier.

Low-pass LC audio filters are common although simple resistor-capacitor combinations are sometimes found. Surplus 88-mH telephone toroid coils are often used in the audio filter, **fig. 9**. This arrangement uses an integrated circuit as a product detector. The RCA CA3028A integrated circuit is a simple differential amplifier with a means of applying an oscillator signal to the base of the constant-current emitter-bias circuit of the differential pair (**fig. 10**). Note in **fig. 9** that the incoming signal is applied between terminals **1** and 5; the oscillator to terminal 7. Demodulated audio is taken from pins 6 and 8.

An fet product detector[10] is shown in **fig. 11**. The input signal is applied to the source; oscillator, to the gate. A simple resistor-capacitor filter passes the audio

products of the drain circuit to the fet audio-input state.

Probably the simplest product detector is that used by PAØKSB,[11] fig. 12. A two-diode balanced mixer is followed

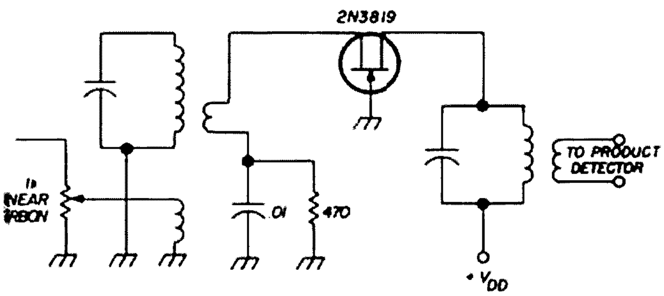


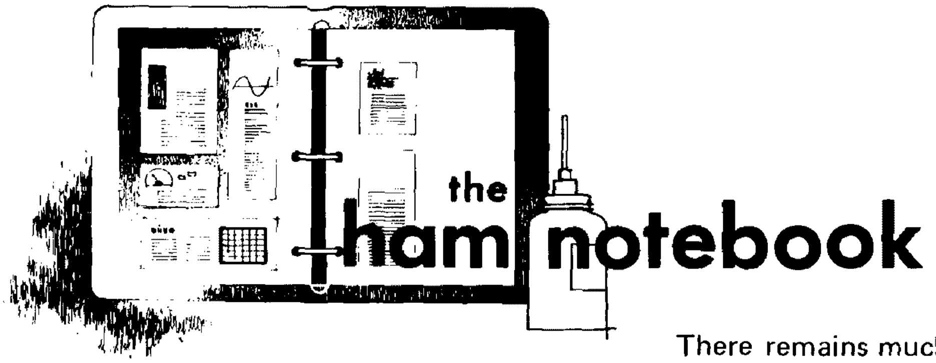
fig. 13. G3EJF's common-gate fet rf amplifier for use with direct-conversion receivers.

by a resistor-capacitor audio filter and a two-stage audio amplifier for headset operation. For these simple detectors a good low-noise fet input rf stage is a definite benefit. A grounded-gate stage from G3EJF[7] is shown in fig. 13.

references

1. Leo Gunther, VK7RG, "State of the Art," *Australian EEB,* April, 1971, page 32.
2. C. F. Dorey, "A Direct-Conversion Receiver for 14 MHz," *Radio Communication,* May, 1970, page 296.
3. Doug Demaw, W1CER, "The DC 80-10 Receiver," *QST,* May, 1969, page 11.
4. Pat Hawker, G3VA, "High-Performance Direct-Conversion Receivers," *Radio Communication,* November, 1970, page 760.
5. Richard S. Taylor, W1DAX, "A Direct-Conversion SSB Receiver," *QST,* September, 1969, page 11.
6. Pat Hawker, G3VA, "Synchrodyne Transceiver Ideas," *Radio Communication,* September, 1970, page 603.
7. J. E. Hodgkins, G3EJF, "A Portable CW Transceiver for 3.5 MHz," *Radio Communication,* November, 1970, page 757.
8. Doug Demaw, W1CER, "Once More With QRP," *QST,* August, 1970, page 17.
9. Wes Hayward, W7ZOL, and Dick Bingham, W7WKR, "Direct Conversion, a Neglected Technique," *QST,* November, 1968, page 15.
10. Rick Littlefield, K1BQT/6, "The Sideband Minituner," *ham radio,* October, 1970, page 16.
11. Pat Hawker, G3VA, "PAØKSB's Two-Phase D-C Converter," *Radio Communication,* November, 1970, page 761.
12. Richard S. Taylor, W1DAX, "CW Transceiver Operation with Transmit-Receiver Offset," *ham radio,* September, 1970, page 56.

ham radio



component marking

There are several types of typing correction tape that are ideally suited for marking printed-circuit boards, transistors and other components. I use the type that is about 1-inch wide and comes in a dispenser (Sterling Type-White or Dixon Taperaser). Place the correction tape over the surface to be marked and letter directly on the back of the tape. For best results use a sharp pencil or ball-point pen. The white correction-tape pigment is transferred easily and legibly.

Because of the stark whiteness of the pigment, readability is excellent, even on light-colored printed-circuit boards. To protect the letters dab a little lacquer or synthetic varnish over the lettering. The white pigment is not especially solvent in lacquer-based materials, but it does have a tendency to smear a little.

W. H. Fishback, W1JE

calibrator crystals

In recent years there have been many items published on the subject of improving the accuracy and reliability of receiver crystal calibrators. Some hams have provided external transistor oscillators, thus tending to limit the temperature range of the unit compared with one exposed to tube heat. Others have gone to external oven-controlled crystals. There remains much interest in improving the internal calibrator, hopefully without modifying the receiver.

The performance of the trimming capacitor in the receiver can be improved by adding parallel and series N750 and npo capacitors to restrict the trimmer range and to obtain better temperature compensation.

There has been little comment, if any, upon the simple approach – replacing the internal calibrator crystal with a better one.

Herb Blasier, W6EF, points out that the usual receiver has an inexpensive E-cut 100-kHz crystal, but this cut has a very small range of temperatures for which it has minimum drift. This crystal may be replaced with a DT-cut crystal for 100 kHz, or for 1 MHz. The DT cut has a long, flat curve, and is little affected by reasonable changes in operating temperatures. For 50 kHz, a useful frequency for calibrators, the NT cut is desirable.

Bill Conklin, K6KA

low-value voltage source

Although zener diodes are available in a wide veriety of voltages, values below about three volts are not as common as other values. Such regulating voltages are sometimes needed and can be built from only a few discrete components. The circuit in **fig. 1B** is a transistorized equivalent of a zener diode. The breakdown

voltage is adjusted by resistor R2. This circuit has been used to obtain regulating voltages from less than 1 volt up to 10 volts.

Circuit operation is quite simple. When the voltage across R1 reaches approximately 0.5 volts, Q1 begins to conduct, causing Q2 to conduct. Q2 draws only enough current through load resistor R_L to maintain conduction in both Q1 and Q2. This voltage is the breakdown voltage V_Z. If the supply voltage is increased Q2 will conduct more current but V_Z will be maintained.

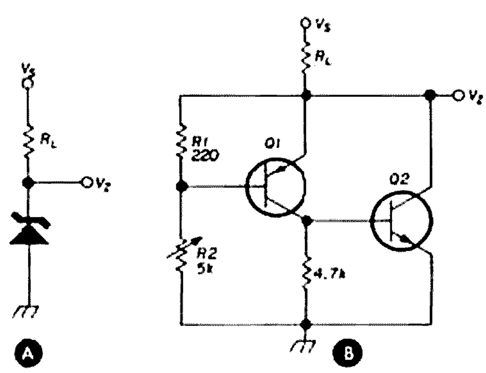


fig. 1. Circuit in (B) is transistor equivalent of circuit in (A). Output voltage can be adjusted from less than 1 volt up to 10 volts.

The regulating voltage, V_Z, is the equal to:

$$V_Z = 0.5\left(\frac{R1 + R2}{R1}\right)$$

The value of the load resistor R_L is determined as in the conventional zener-diode circuit in fig. 1A:

$$R_L = \frac{V_S - V_Z}{I}$$

where I is the current through zener and the load.

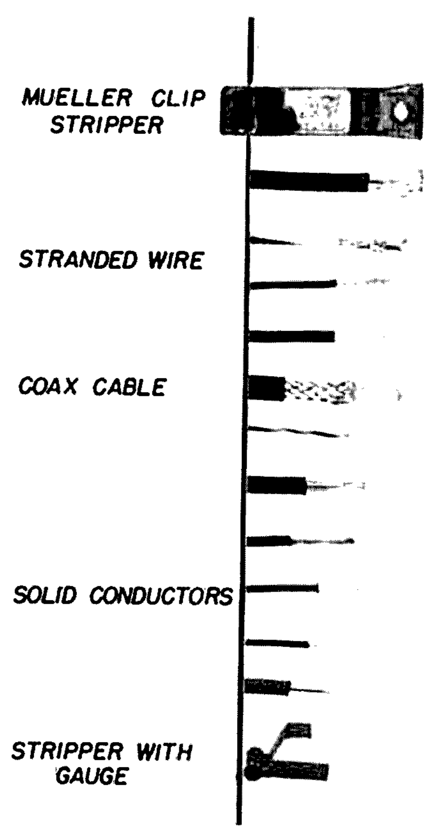
This circuit can be used in almost any application that a regular zener diode can be used, but it has one advantage that can often be very useful to the experimenter — the regulating voltage can be adjusted to precisely the required value.

Transistors Q1 and Q2 should be low-leakage silicon transistors. The power rating of Q2 determines the power rating of the equivalent zener. Another advantage now becomes apparent: an expensive high-power zener diode can be replaced if an appropriate power transistor (low cost) is used for Q2!

James McAlister, WA5EKA

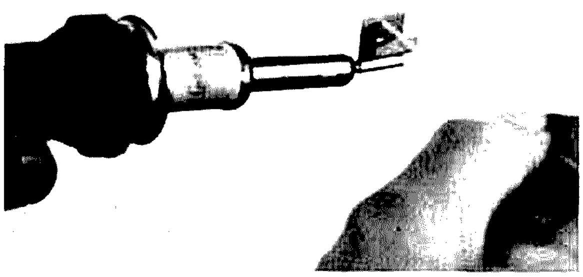
hot wire stripper

The wire strippers shown in the photographs permit consistently better stripped wires when building electronic equipment. I use two different versions with great success. Both use a hot knife edge to melt the insulation in a thin line around the wire, permitting clean removal of insulation without damage to the conductor. Cut strands or nicks are completely eliminated. The stripper can be used on various sizes of solid conductors, stranded wire or coaxial cable. It is especially useful on very small gauge stranded wire. On coaxial cable clean cuts can be made through the outer jacket and, after moving the braid out of the way, through the insulation around the center conductor.



Hot-wire strippers and samples of stripped wire and cable.

For the first version of the hot wire stripper remove the screw from a Mueller battery clip, flatten this end and file to a dull knife edge. Clip to the tip of a soldering iron (100 watts or less) and when the edge is hot enough to melt the plastic insulation the stripper is ready for use. Place the knife edge against the insulation and rotate the wire between your fingers until a clean cut is made to the conductor. The insulation will slide off cleanly.



Wire stripper with gauge installed on 37-watt Ungar soldering iron.

The second version, for use with a 37-watt Ungar soldering iron, has a built-in stop to gauge the length of insulation to be stripped off the wire. The dimensions are not critical, and materials used are available from scrap. The stripper shown in the photograph was made from a steel metal strap used for banding cardboard cartons. It is .010-inch thick and ½ inch wide, bent around a nail to the shape shown in the photograph. Both ends are filed to a dull knife edge. The stop should extend about 1/8-inch beyond the other knife edge to permit insulation of any length to be stripped. To use this feature place the end of the wire against the stop and rotate the wire between your fingers until a clean cut is made. When stripping thick insulation or on coax cable, using either stripper, roll the cable on your bench with a small amount of pressure on the stripper.

Carl Yeager, W8DWT

cleaning tape heads

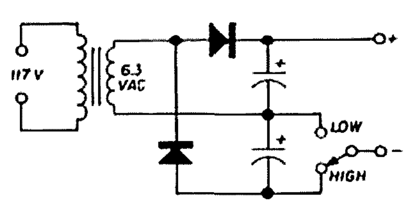
Sound recorders are used widely by amateurs, many of whom are not experienced audiophiles. After awhile, a recorder may fail to record at a suitably high level until the recording and erase heads are cleaned. In my case, I had purchased a bottle of cleaner, and also tried rubbing alcohol which did not correct the loss of recording volume.

One salesman said later that it was necessary to use an almost pure alcohol, not a 70% rubbing alcohol. The instruction book recommended using carbon tetrachloride, which has some serious toxic effects if much is used in a closed space and is impossible to buy in some states. However, in checking with users of commercial recorders and computers, it was found that Ampex No. 087-007 head cleaner consists of another somewhat toxic material, less than 11 percent Trichlorethane. A little of this on a Q-tip quickly corrected the recording volume.

Bill Conklin, K6KA

dual-voltage power supply

Many home construction projects involving the use of transistors call for a 9-volt power supply. Once in a while you may want a bit higher voltage, say 18 volts. Here's a circuit for getting either of those two voltages from a power supply using a transformer with a 6.3-volt ac

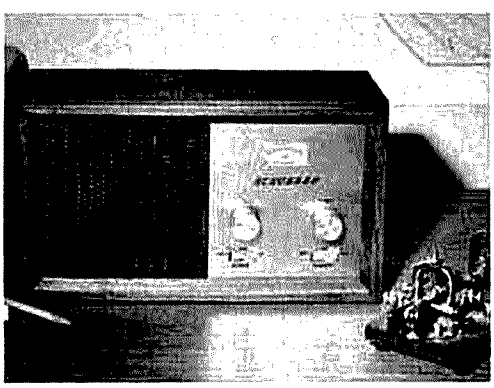


secondary, the size that lives in almost every junkbox. The wiring diagram in **fig. 2** tells about all that can be said. The transformer, rectifiers and capacitors are those that are available. The spdt switch selects either of the two available voltages.

Carl Drumeller, W5JJ

new products

power supplies



The *Scrubber* is a solid-state aid to noise-free CW reception. It also serves as a station speaker and code practice oscillator. The *Scrubber* contains a *Fritch*,* a sharp, solid-state, active filter which "scrubs" the audio output from the receiver. The filtered signal is rectified and used to drive a dry-reed relay. The relay contacts are used to key an internal audio oscillator, whose tone and volume are selected by the operator. The code output matches the received signal but is isolated and free of noise.

The filtering and oscillator keying functions are performed by the Douglas Randall *Fritch* frequency responsive switch. The *Fritch* provides a contact closure when a 1000-Hz signal is received from the receiver audio section; the contact closure is used to key the *Scrubber* oscillator.

The *Scrubber* contains a matching transformer input, *Fritch,* monitor oscillator, communications quality speaker and power supply. A rear panel terminal strip provides connections for receiver audio and practice key. Provisions are made to match 3.2-, 8-, 600- and 1000-ohm receiver outputs. A rear panel adjustment sets the proper signal input level to the *Scrubber.*

In operation, a CW signal is tuned in with the *scrub* switch *out*. When the desired signal peaks on the front panel meter, the *scrub* switch is thrown to *in.* The desired audio signal will then key the internal oscillator in step with the received signal. The extremely high degree of isolation between the actual audio signal and the regenerated signal eliminates all unwanted static and noise. The *Scrubber* will pass signals in a very narrow audio passband and reject all others.

Like any filter, the *Scrubber* cannot distinguish between multiple signals in its passband. However, rough or chirpy signals are reproduced as clear T9 notes. The better the receiver selectivity, the easier it is for the *Scrubber* to separate signals. SWLs and novices will find that the *Scrubber* will improve copy from homebrew and less expensive receivers and will eliminate distracting QRN. Ear fatigue from static crashes is eliminated.

The wide dynamic input range of the *Scrubber* helps to compensate for fading signals. The use of a reed relay allows faithful reproduction of the original signal at speeds in excess of 25 words per minute.

The Scrubber is housed in a handsome wood-grained cabinet and can also be used as a station speaker, CW monitor or code-practice oscillator. The unit is priced at $94.50. For more information, use *check-off* on page 110, or write to Douglas Randall Division, Walter Kidde and Company, Inc., 6 Pawcatuck Avenue, Pawcatuck, Connecticut 02891.

*Fritch is the registered trade mark for Douglas Randall frequency-responsive switch; patent pending.

More Details? CHECK—OFF Page 110

printed-circuit kit

At one time building an electronics project was hard work because it involved drilling, cutting, reaming and deburring a metal chassis. Now Injectorall Electronics Corporation has taken the sweat out of electronics and put pleasure back in. With Injectorall's new no. 650 photo-etch printed-circuit kit, anyone, beginner or professional, can make professional printed circuits the first time, every time. Injectorall's 650 is a completely packaged kit (nothing else to buy) using a photosensitive method for producing professional quality printed circuits. It can be used with assurance by engineers developing a prototype or an amateur who is building a home-lab project.

With Injectorall's Kit 650 you don't need a darkroom, and you can completely eliminate commercially made boards. If *ham radio* has a drawing you want to use, Kit 650 has materials included to let you make negatives from the printed page. Hobbyists and professionals alike have found it ideal for solid-state and integrated circuitry.

Kit 650 contains two photo-sensitized 3 x 4-inch copper-clad boards, a photographic test negative and an ultraviolet light source. It also contains an exposure glass, clamps, developer, etchant, trays, resist remover, drill and complete instructions. This low-cost easy way of making quality printed circuit boards is now available at all major distributors and retails for only $10.80.

For more information, use *check-off* on page 110, or write to Injectorall Electronics Corporation, 4 North Road, Great Neck, New York 11024.

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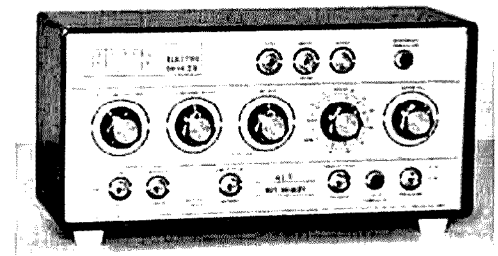
programmable electronic keyer



Curtis Electro Devices has announced an unusual new electronic keyer designed to raise contest scores while adding to the enjoyment of CW operation. A 256-bit MOS integrated-circuit read-write memory has been incorporated into the basic Curtis keying circuitry to provide "canned" CW sequences for repetitious portions of CW traffic such as used during CW contests. This frees the operator for logging and provides a moment of relaxation. The following is a typical Field Day program set:

A. 5NN SCV DE W1DTY/1
B. CQ FD CQ FD DE W1DTY/1
C. DE W1DTY/1

These three programs are *written* into the MOS memory merely *by sending* on the keyer paddle while the instrument is in the *write* mode. The sequence may be written and rewritten as often as desired for any contest or call letters in only the time it takes to send the Morse characters. An internal battery automatically supplies current to the memory in the event power is removed. The memory may be programmed at any convenient speed. Playback, initiated either by a panel push-button or an external switch, exactly follows the manual speed and weight characteristics in use by the operator and is indistinguishable from manual sending. Memory reset is instantaneous when manual sending is started.

The all IC EK-402 Programmable Electronic Keyer incorporates the features required in a professional keyer including calibrated speed, variable character weighting, sidetone pitch and volume, a variable repeat cycle, tune-up switch,

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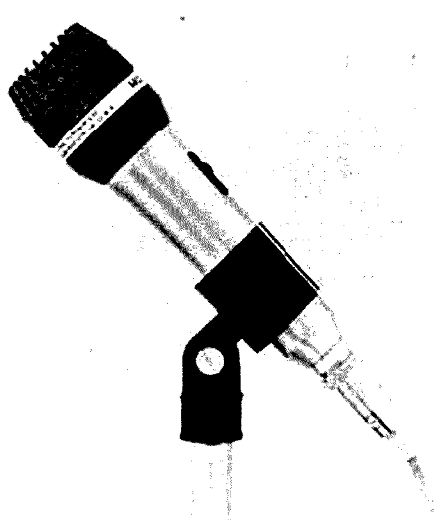
iambic or standard keying action and dot memory. Speaker and 110 Vac power supply (220 Vac optional) are built-in. The reed relay output will key any grid-block, cathode keyed or solid-state amateur transmitter. The keyer is priced at $289.95 FOB factory, complete with cables, connectors and memory battery. For more information write Curtis Electro Devices, Box 4090, Mountain View, California 94940, or use *check-off* on page 110.

turner microphones



The medium-priced Turner 600 cardioid dynamic microphones now include a detachable cable with professional connector, and the guarantee has been extended to five years at no increase in price. The Turner 600 series ($70 list) has excellent unidirectional cardioid pick-up characteristics and a smooth wide-range response. The units feature die-cast zinc alloy cases, Cycolac fronts, on-off switches and are available as a Model 600 high impedance (40,000 ohm) or Model 602 low impedance (150 ohm). Response covers the 50-15,000 Hz range with discrimination typically 20-25 dB over the entire range.

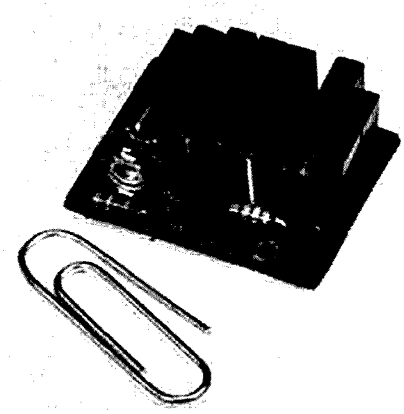
Effective immediately, the permanently attached cord on the 600 Series microphones is replaced by a 20-foot detachable cord, including an Amphenol MC2M connector. The new cable is single-conductor, shielded, on high-

impedance models, and two-conductor, shielded, on low-impedance models.

The guarantee on the 600 series microphones has now been extended to five years, longest in the microphone industry. If, at any time within five years of purchase, the microphone fails, it may be returned to Turner, prepaid. If the failure is due to faulty materials or workmanship, the microphone will be repaired or replaced, at no charge, and returned to the sender, postage paid.

The Turner Division of Conrac Corporation, located at 909 17th Street N. E., Cedar Rapids, Iowa 52402, is one of the world's leading manufacturers of microphones for broadcast, sound systems, amateur radio and other applications. For more information use *check-off* on page 110.

sub-audible tone module



Alpha Electronic Services has announced its new SS-80J integrated-circuit sub-audible tone module. Measuring less than one square inch, including the new TN-91J frequency determining module, the SS-80J is the smallest non-reed CTCSS device available. The unit was designed especially to offer high reliability by the elimination of mechanical and contactless reeds and still be small enough to be installed in hand-held transceivers and pocket page receivers.

Under development for two years, the SS-80J has been subjected to a rigorous

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testing program to establish long-term reliability. Extremely stable, the unit meets or exceeds all applicable EIA Specifications and is available in standard or special frequencies from 20.0 cps to 250.0 cps. Special configurations are available for easy mounting in most hand-held and portable units along with instruction manuals that clearly outline step-by-step installation. For additional information write to Alpha Electronic Services Inc., 8431 Monroe Avenue, Stanton, California 90680, or use *check-off* on page 110.

instrument knobs

International Rectifier Corporation has introduced a family of all aluminum, quality instrument knobs at prices ranging from 56 cents to $1.19. The knobs are designed specifically to upgrade the appearance of instruments built by hobbyists achieving a custom, precision look.

In the International Rectifier Diamond line, the company will offer three styles of individually machined knobs in a total of 11 sizes. These include a *silver* aluminum knob style with knurled sides and recessed spin top. This style is also available in skirted configuration. The third series is gold anodized aluminum with recessed tops and wood-grained inserts.

All three styles are individually packaged with a 1/8-inch adapter to make the standard ¼-inch shaft fit all sizes. Each knob has a set screw to assure secure fit. Knob sizes range from approximately ½ inch to 1¼ inch in diameter. Available at your local electronics distributor; for more information use *check-off* on page 110.

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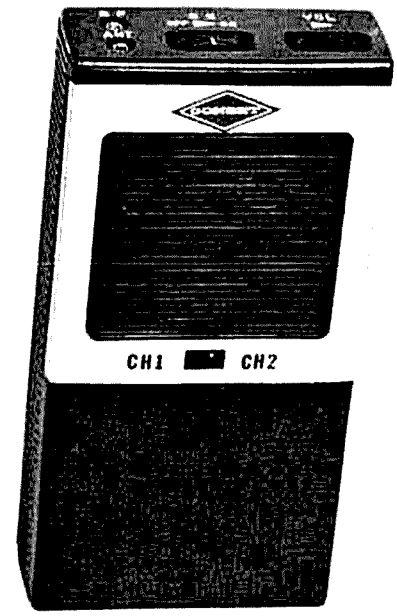
gonset portable vhf receiver



The Gonset Division of Aerotron recently announced its new model 6RP highband vhf personal portable receiver. The Gonset 6RP is a high-performance, true fm narrow-band pocket-type receiver, with built-in antenna for monitoring the range of 146 to 160 MHz. The 6RP may be powered with replaceable mercury cells or rechargeable nickel cadmium batteries; other options include two-frequency capability, and signal-to-noise ratio squelch which is adjustable from the outside of the unit. The Gonset 6RP may be used for monitoring amateur fm, police, fire, industrial, CD and many other land-mobile services.

For more information on the new Gonset 6RP vhf-fm receiver, use *check-off* on page 110 or write to Aerotron, Inc., Post Office Box 6527, Raleigh, North Carolina 27608.

second-class radio-telephone license handbook

The emphasis of this new edition, authored by Edward M. Noll, is placed on two-way radio since this is the major field of activity for the second-class radio-telephone license holder. Completely updated, the discussions of the latest solid-state two-way communications

 More Details? CHECK—OFF Page 110

equipment are included in this revised edition.

The first five chapters cover FCC rules and regulations, operating procedures for radiotelephone communications, transmitter tuning and adjustment, etc. Chapters 6 through 14 are study guide chapters that contain questions and answers on Elements I, II and III of the FCC exam. Each of the study guide chapters include a self-examination, with the answers provided.

Chapter 15 contains two 100-question tests to simulate the FCC examinations. These tests are presented in the multiple-choice form of the FCC tests. The answers to these tests are also given. Finally, the book has appendices which contain extracts from the FCC Rules and Regulations. These extracts contain valuable information that can be used as reference material.

Second-Class Radiotelephone License Handbook will help the reader acquire a license, and will continue to serve as reference material once he is licensed. Published by Howard W. Sams & Company. 360 pages, softcover. $6.50 from Comtec Books, Post Office Box 592, Amherst, New Hampshire 03031.

vhf fm transmitter kit

The new two-meter five-watt vhf fm transmitter kit from RMV Electronics features 4-channel capability. The kit is furnished with step-by-step instructions and is built on a double-sided 3- x 7-inch glass-epoxy board. The circuit uses 16 transistors and 9 diodes and has a separate oscillator for each channel, capable of remote selection. The transmitter uses plug-in HC-18/U crystals in the 9-MHz range.

The completed transmitter is easy to tune up with a 50 μA meter and provides 5 watts (typical) output with a 13.6-volt power supply. Priced at $59.95 less crystals, microphone, meter or power source from RMV Electronics, Box 283, Wood Dale, Illinois 60191; add $1.00 for postage in the U. S. A. For more information use *check-off* on page 110.

More Details? CHECK—OFF Page 110

75 cents

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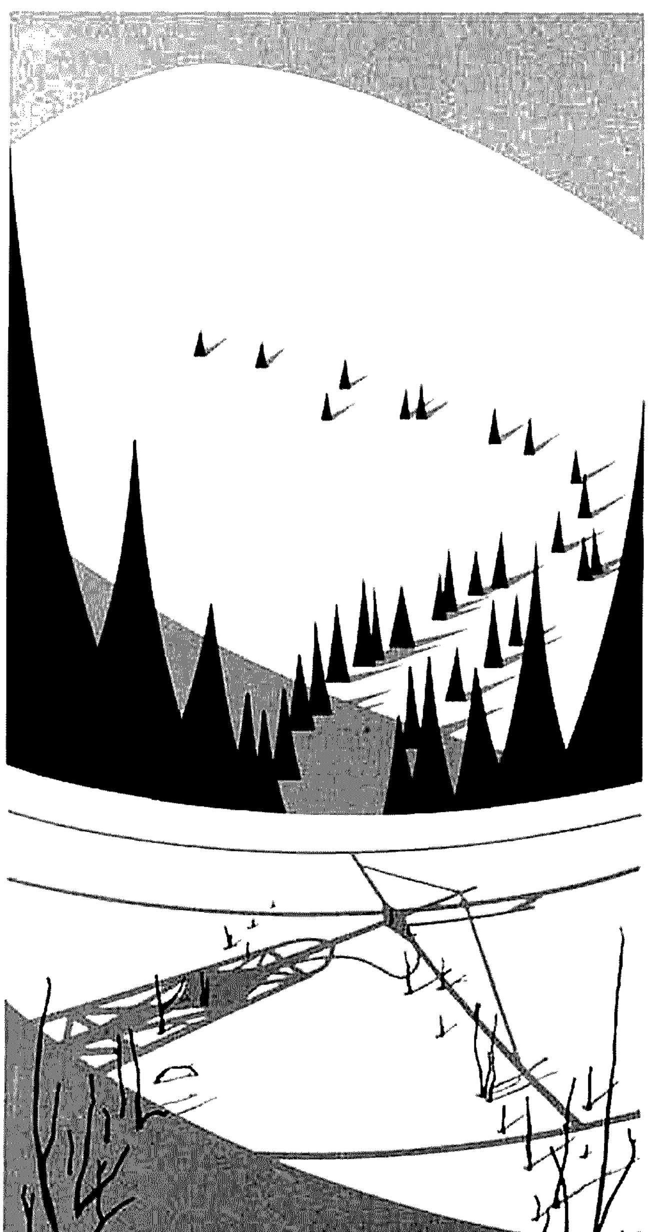
focus on communications technology . . .

ham radio magazine

DECEMBER 1971



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ham radio magazine

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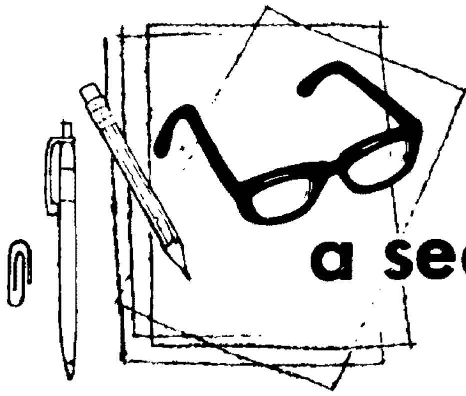


a second look by jim fisk

Radio engineers from the U. S. Department of Commerce are creating immense, invisible heated bubbles in the upper atmosphere with a new 100-million-watt radio beam. The short-lived bubbles are formed in seconds at altitudes up to 200 miles and grow to their full 50 to 100-mile size in about 20 minutes. Composed of the electrified gas of the ionosphere, the bubbles elongate upward and downward under the force of the earth's magnetic field.

Observations of these effects on the upper atmosphere should lead to a better understanding of the ionosphere and improvements in long-distance radio communications. Within minutes or hours, depending on the time of day and conditions in the high-level environment, the modified region of the ionosphere rebounds to its natural state.

The new high-power transmitter, located in Colorado, uses a nine-element circular array of antennas, 360 feet in diameter, with an additional element in the center for beam focussing. The system is capable of projecting effectively a 100-megawatt radio beam that is tunable between 5 and 10 MHz, the usual range of ionospheric penetration frequencies.

The intense radio beam is transmitted straight up at very close to the penetration frequency — the frequency at which a radio wave passes completely through the ionosphere. This imparts a maximum amount of energy to the ionosphere. The closer the transmitted beam approaches the penetration frequency, the higher it reaches before being bent back to the earth, the more it is slowed down, the longer it remains in the ionosphere, and the more its energy is absorbed by the ionospheric electrons.

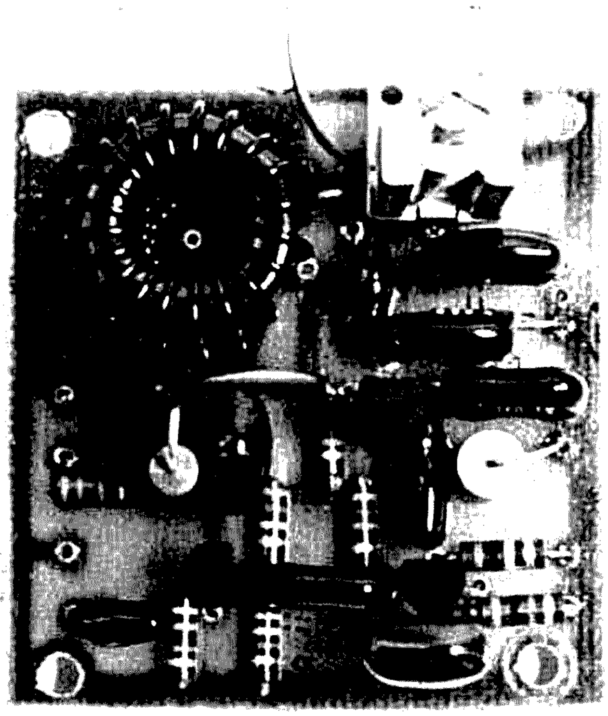
The electron heating takes place in about 20 seconds, more or less. The heat bubble expands more slowly, however, because the negatively-charged electrons must drag the heavier, slower-moving positive ions with them in order to maintain the electrical neutrality of the ionospheric plasma. Within 20 minutes or so the dimensions of the heat bubble may grow to 50 or 100 miles.

Measurements indicate that the radio beam raises the temperatures of the ionospheric electrons by as much as 35 percent. Scientists expected that this temperature change would cause enhanced reflections of radio signals. Unfortunately, this is not the case; radio waves reflected from the regions of heated electrons are severely attenuated.

Another major surprise is the artificial creation of a natural phenomenon known as Spread-F. Spread-F is the upper layer of the radio-reflecting region of the ionosphere which is characterized by a patchy pattern of reflected signals.

Scientists have long been looking for a technique for modifying the ionosphere as a method for studying it. They have used chemical releases, atomic bombs and small electron-beam accelerators, but none of these methods are as controllable or repeatable as the new 100-megawatt transmitter. Hopefully this new tool will lead to a better understanding of the ionosphere, and eventually, control of its radio reflecting properties. Can you imagine around-the-clock contacts on 10, 15 and 20 meters even in years of minimum sunspots? QRM could become more of a limiting factor than it is now!

Jim Fisk, W1DTY
editor



miniature solid-state variable-frequency oscillator

This miniature circuit-board can be used for either the Seiler or Vackar oscillator circuits, including the buffer-amplifier and emitter-follower

Donald R. Nesbitt, K4BGF

My search for a simple high-quality vfo which was small enough to justify the use of minature solid-state devices led to the design of a dual-function printed-circuit board which accommodates either the Seiler or Vackar oscillator circuits. These circuits were given comprehensive treatment by W1DTY[1] and can provide the amateur with exceptional results.

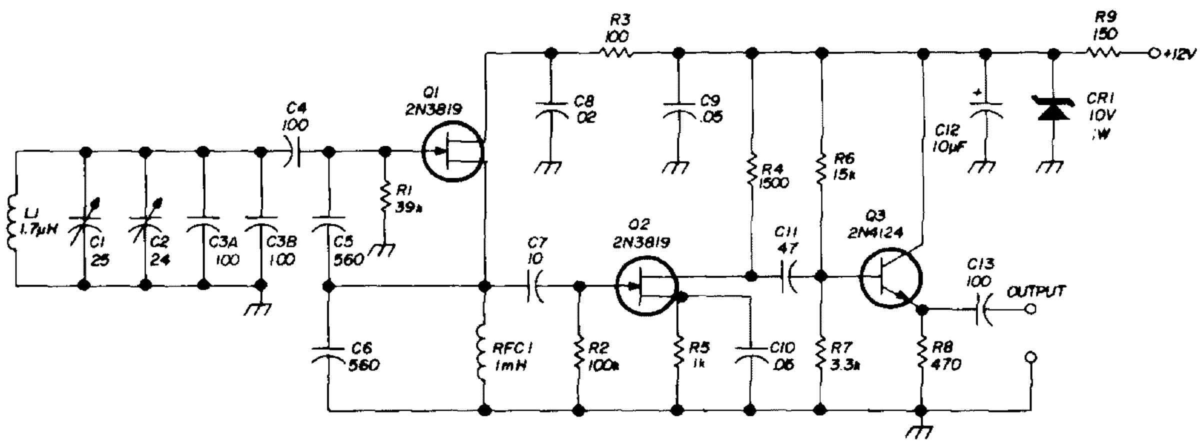
seiler and vackar circuits

Fig. 1 shows the Seiler oscillator circuit. Q1 is the oscillator, followed by an amplifier (Q2) and an emitter-follower (Q3). **Fig. 2** shows the schematic diagram of the Vackar. Again, Q1 is the oscillator, followed by an amplifier (Q2) and an emitter-follower (Q3) which are identical

to those used in the Seiler circuit.

Careful study of the basic Seiler and Vackar oscillator circuitry reveals that the circuit differences involve a basic rearrangement of L1, C6, C7 and RFC1, and, in the case of the Vackar, the elimination of C8 and R3.

The padding capacitor, C3, is designated as C3A and C3B. This allows you to use two different temperature compensating values should this be necessary.



C1	25-pF variable (Millen MK21025)
C2	2.4-24.5 pF (Johnson 189-509-5)
C3	200 pF silver mica (total of C3A and C3B)
L1	1.7 µH. 18 turns no. 20 enameled evenly spaced on Amidon T-68-2 toroid core

fig. 1. Circuit of the Seiler oscillator. Components values shown tune from 7.0 to 7.3 MHz.

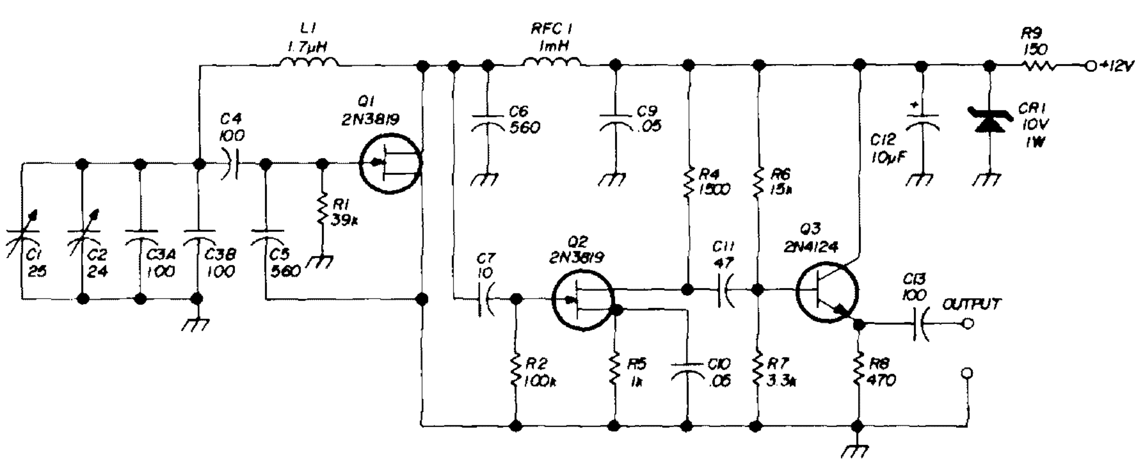


fig. 2. Vackar oscillator circuit. Buffer-amplifier (Q2) and emitter-follower (Q3) stages are the same as the Seiler circuit in fig. 1.

circuit board

Fig. 3 shows the parts placement as viewed from the foil side of the printed-circuit board for *both* the Seiler and the Vackar circuits.* Connections shown by dashed lines are for the Vackar; its associated components are designated by the letter V (ie: V-L1, V-Jumper, V-C6, V-C7, and V-RFC1).

Parts placement for the Seiler circuit can be seen by simply ignoring the

*An etched and drilled glass board (including the nine quarter-watt resistors used in the test oscillator) is available for $1.50 postpaid ($1.00 for the board only) from Technical Assistance Unlimited, Inc., 717 N. W. 1st St., Gainesville, Florida 32601.

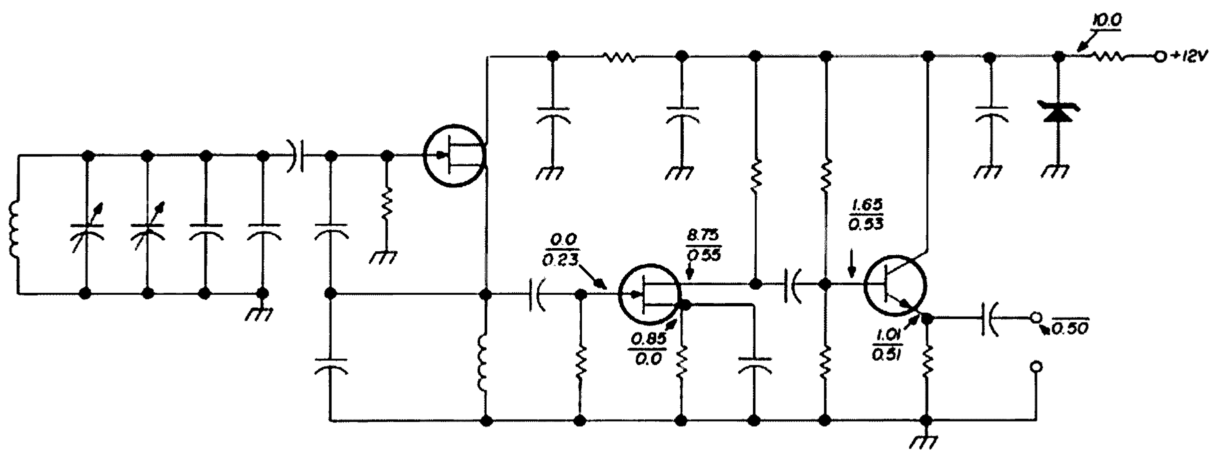


fig. 4. Dc and rf voltage measurements made on the 7-MHz Seiler test oscillator. The dc voltage measurement is the upper number; the rf voltage (rms) is below the line.

dashed lines and the V-designated connections. When the circuit board is used for the Seiler circuit *no* connections are made to holes indicated by the letters E or F.

If you begin with the Seiler oscillator,

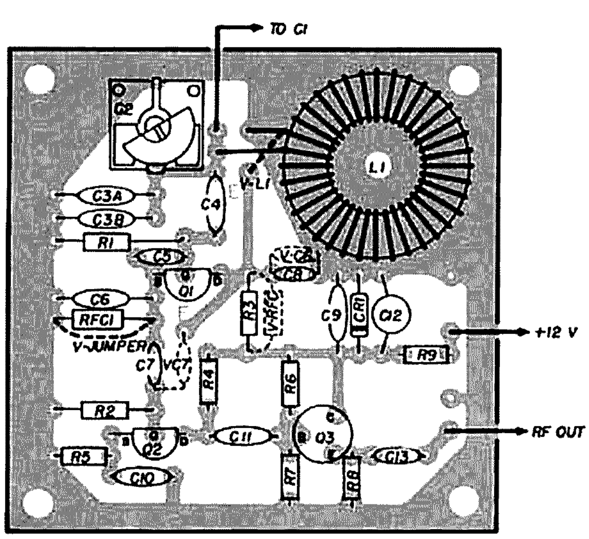


fig. 3. Parts placement as viewed from the foil side of the printed-circuit board. Seiler layout can be seen by ignoring the dashed lines and components designated by "V".

the following changes are necessary to convert to the Vackar circuit:

1. The side of L1 which was connected to ground is connected to point E.
2. RFC1 is replaced by a jumper.
3. R3 is replaced by RFC1.
4. C8 is replaced by C6.
5. The end of C7 which was connected to the source of Q1 is connected to point F.

After these five steps are completed, C8 and R3 are leftover as they are not required in the Vackar configuration.

dc and rf voltages

Fig. 4 indicates the dc and rf voltage (rms) measurements made on a 7-MHz Seiler test oscillator. Circuit values are shown in fig. 1. These measurements were made with an 11-megohm vtvm equipped with an rf probe. The upper number indicates the dc voltage with Q1 removed from the circuit. The bottom number is the rf voltage (rms) obtained with the oscillator functioning.

construction and test

Since this circuit board packs a lot of components in a small space it is recommended that you mount the resistors first. The quarter-watt size is preferred but half-watters may be used by standing them on end. Next, mount the capacitors, coil, choke, and zener — checking and re-checking for proper position.

The transistors are mounted last. You should mount Q2 and Q3 and check the dc voltages as shown in fig. 4. Any abnormal reading will allow you to find

the mislocated component or faulty transistor. Install Q1, apply power and you should be in business.

The rf voltage measurements shown in **fig. 4** for the 7-MHz test oscillator can be used as a rough guide. Using the Vackar circuit, or other frequencies or components will result in rf voltages that depart widely from those of the test oscillator. Once you have obtained proper operation, you may wish to note these voltages for reference.

Failure to oscillate may be traced to a mislocated part, a bad fet or improper values at C4, C5, C6 or L1 and its associated parallel capacitors. In many cases you may have to experiment with these components to determine the proper value. In addition to the comprehensive article by W1DTY already mentioned, the additional references[2, 3] provide examples of various frequency ranges and circuit layouts.

Generally speaking, mechanical construction of the finished product will spell the difference between a "warbler" and crystal-controlled stability. The general rule of thumb is to build it like the venerable battleship; the article by W8YFB[4] is recommended reading.

"You should take up ham radio too, Marge! It's a great way to while away the idle hours."

By all means try your hand at substitutions. Some substitutions may produce superior performance. MPF102 fets may be used in place of the 2N3819s but be sure to check the base lead configuration as the gate lead is on the outside instead of in the middle. Some bending may be necessary to get the right lead into the right position, and care should be taken to insure that none of the leads are short-circuited. This substitution was made in the test oscillator, and it functioned with little perceptible difference from the original except for slightly higher rf voltages.

Toroidal inductors need not be used at L1 although they do offer a size advantage. Miniductor type stock was tried and performed very well. Alternately, the hole used to mount the toroid can be enlarged and a slug tuned coil fastened in this position. While this works, it may be necessary to adjust some of the components to get the circuit oscillating as the Q of this type inductor may be considerably lower than either the toroid or the airwound coil.

Various rf chokes with values as low as a few microhenrys were tried at RFC1 with success.

summary

This versatile printed-circuit board permits easy construction of either the Seiler or Vackar oscillator. In addition to being useful to the homebrewer in search of a small high-quality vfo, this board should keep the Seiler vs Vackar experimenter busy for some time.

references

1. James R. Fisk, W1DTY, "Stable Transistor VFOs," *ham radio,* June, 1968, page 14.
2. James R. Fisk, W1DTY, "Homebrew Single-Band SSB Transceiver," *ham radio,* June, 1969, page 11.
3. C. E. Galbreath, W3QBO, "A VFO For Solid State Transmitters," *ham radio,* August, 1970, page 36.
4. Bill Wildenhein, W8YFB, "Simple High-Stability Variable Frequency Oscillator," *ham radio,* March, 1969, page 14.

ham radio

diversity receiving system

The application of photocell modules and ICs to improve reception in the hf bands

John J. Schultz, W2EEY, 1829 Cornelia Street, Brooklyn, New York 11227

In the same sense that most family garages these days contain two or more cars, most amateur stations have at least two receivers. The second receiver may be a separate self-contained unit, or of the two receivers, one may be in the form of a transceiver. The availability of two receivers covering the same band or bands allows one to enjoy the advantages of diversity reception.[1]

The most useful application of diversity reception in amateur work is probably for ssb DX, since deep signal fading often causes missing words or entire phrases. Although diversity reception will also improve CW intelligibility under marginal signal conditions, the improvement is not quite as obvious as that with ssb since CW speed versus fade time, plus the fact that CW contacts often follow a standard format, allow one automatically to fill in missed portions. Much the same is true of RTTY, where one can see the printout of several words and fill in a garbled word.

Although commercial and government stations may use racks of equipment to obtain diversity reception with low error rates on hf circuits, the basic advantages of diversity reception can be obtained for amateur purposes with relatively simple equipment.

diversity reception

Besides having two receivers available, the other basic requirement for diversity reception is that each receiver be fed an input signal that doesn't fade in the same manner, so that the diversity circuits can choose automatically the best receiver channel at any instant. There are numerous ways to provide the differing input signals for each receiver. The commercial/military approach is usually to use space diversity. In this method, each receiver is fed by a similar antenna separated 5 to 10 wavelengths. The space between the antennas provides different signal levels as ionospheric refraction conditions vary.

Another approach is frequency diversity in which only one antenna is used, but each receiver is tuned to a different frequency which carries the same intelligence. This method is satisfactory for shortwave broadcast reception, since most major shortwave broadcasters transmit similar programs simultaneously on at least two bands to a given area. Amateurs,

however, may transmit simultaneously on different bands only under restricted conditions generally defined by the FCC as serving the overall amateur interest (code practice) or during emergencies.

polarization diversity

Probably the most satisfactory diversity receiving method for hams is polarization diversity, which doesn't require the real estate necessary for space diversity. Its principle is based on the fact that a signal refracted by the ionosphere experiences a change in polarity. The change *is most pronounced* during disturbances in the ionosphere that produce rapid fading. Most refracted signals exhibit circular polarization, and the best hf DX receiving antenna would probably be one that responds to this type of wave. However, such antennas are generally not practical on hf.

Polarization diversity reception isn't difficult to implement in most cases, since all that's required is an antenna of opposite polarization to complement the existing antenna. For instance, if the existing antenna is horizontally polarized, a second, vertically polarized antenna is needed. The gain of the complementary antenna is not as important as the fact that it responds to oppositely polarized signals.

A rarely used method of diversity reception involves (a) cross-response, and (b) extremely different angle-of-arrival-response antennas. In the former case, similarly polarized antennas are used but have a 90° difference in major-lobe response to account for the fact that signals may be received better over different propagation paths. In the latter case, an antenna ½ wavelength high may be used to complement an antenna 1-2 wavelengths high to take advantage of the fact that, during disturbed propagation conditions, transmission over a long path may vary between 1- and 2-hop F-layer propagation with a corresponding difference in the signal angle of arrival.

Whatever antenna method is used depends upon the construction possibilities at a station. The following parts of the article describe circuits that may be used to obtain an automatic selection of the receiver for best reception assuming a given receiving antenna arrangement.

selection via avc combining

The simplest method to obtain a selective choice of receiver outputs is to tie together the avc bus of two similar receivers, as shown in fig. 1. The receiver with the stronger signal biases the receiver with the weaker signal to reduce its audio output. The receiver with the weaker signal is never completely cut off, however, and produces some noise output. Each receiver is shown with a separate speaker, but often the audio output of each receiver is paralleled and fed through the audio section of just one receiver, or through an auxiliary audio amplifier. Simultaneous tuning of both receivers may be accomplished by using a common local oscillator and beat-frequency oscillator. In this case, the receivers should be adjusted on a steady signal so that both produce the same output level independently. Such an adjustment is usually required only once on each band. This can be done by controlling the gain of any stage not subject to avc, or by controlling the gain of a preamp ahead of one receiver not connected to the avc line. Once adjusted the system will work quite well and will, very roughly, tend to

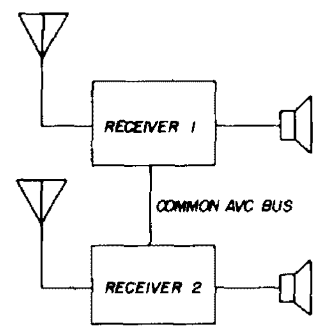


fig. 1. Simplest diversity receiving method. The receiver with the stronger signal biases the receiver with the weaker signal to reduce its audio output. Signal-to-noise ratio suffers, however, because the receiver with the weaker signal is never completely cut off.

halve the difference in input signal level to each receiver. For example, if the input signal to one receiver differs 20 dB with respect to that of the other receiver, the combined audio level will exhibit about a 10-dB change.

inexpensive and performs the audio combining or selection process almost as well as several tube or transistor stages. It's particularly effective when used in receivers with their avc buses tied together. However, it can be used with two re-

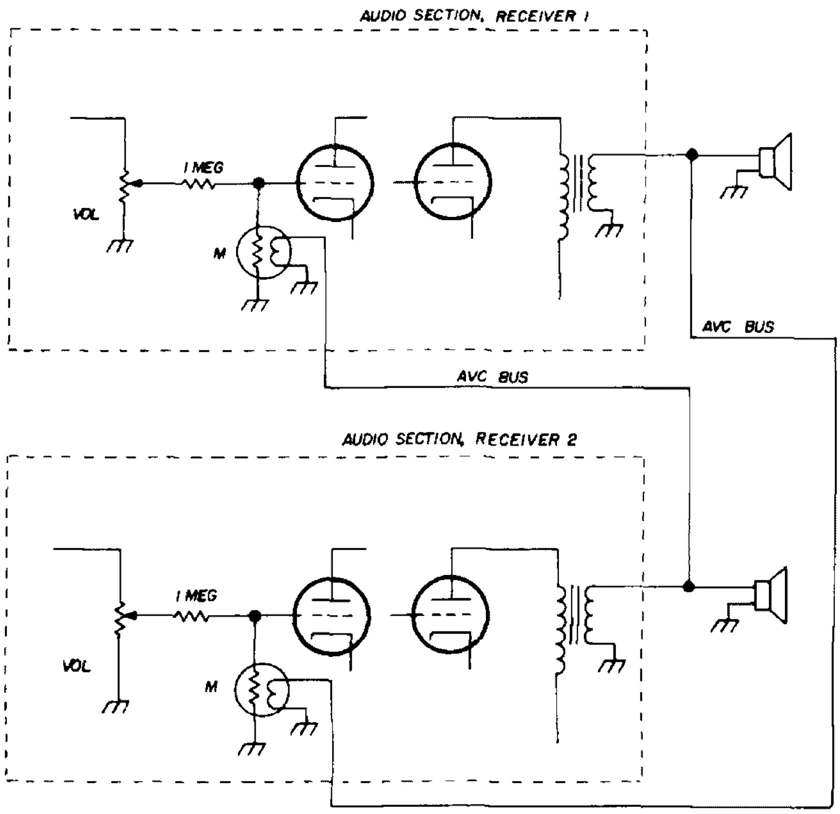


fig. 2. Addition of photocell module, M, improves diversity action of two receivers with avc buses connected, or it can be used alone with two receivers having dissimilar avc circuits. Photocell (Clairex CLM 4006) is available from Allied Radio, order no. 60F6468, at $3.25.

photocell combiner

To further enhance the differential avc voltage when one receiver input signal increases as the other decreases, a photocell operating on the audio output of each receiver can be used to advantage (fig. 2). The receiver with the stronger audio output energizes the photocell module in the first stage of the other receiver. The resistive element of the photocell module decreases in resistance with increasing drive, and in conjunction with the 1-meg series resistor, forms a voltage-divider that reduces the signal to the audio section of the receiver with the weaker signal.

The photocell addition is relatively ceivers having different avc circuits, which precludes connecting their avc buses together. After experimenting with the rf and af gain controls on each receiver, the photocell-only method of audio control performs about as well as the avc-combining method.

switching combiner

The main disadvantage of the simple methods described above is that the receiver with the weaker signal is never completely disabled. The combined avc/photocell method is about as far as one can go with simple circuitry; but even here, unless the difference in signal levels to the receivers is very great, the receiver

with the weaker signal will provide some output and degrade the overall signal-to-noise ratio. The only way to avoid this is to completely disable the output of the receiver with the weaker signal. The rate at which the level of each receiver in a cial equipment. Although diversity action would be enhanced if the unit shown were used with receivers whose avc buses could be tied together, it can be used with receivers having dissimilar avc circuits.

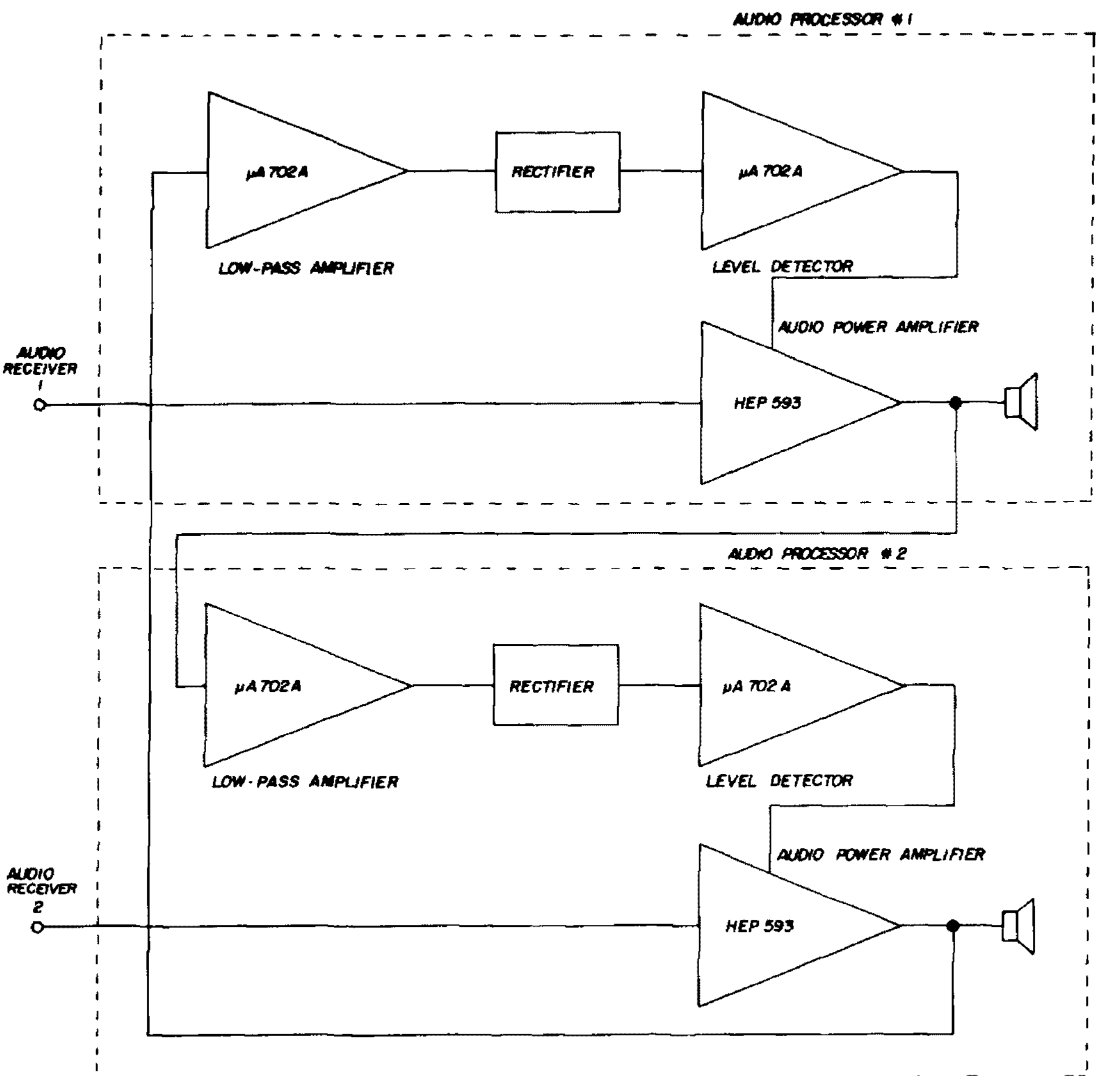


fig. 3. Block diagram of IC diversity combiner. Note that the two audio-processing circuits are identical.

diversity system should be sampled, as well as how fast and by what means the output of each receiver can be chosen without creating switching noise, has been the subject of numerous studies and commercial design approaches.

IC diversity combiner

The simple circuits using ICs shown in fig. 3 and 4 provide effective, low-cost diversity receiving action approaching that obtainable with expensive commer-

As shown in fig. 3, the IC combiner consists of two identical audio processors, each connected to the audio output of one receiver, then interconnected. Each processor consists of a 1-watt audio power amplifier whose output is regulated by a control section consisting of two μA702A stages. The first μA702A stage is driven by the audio-output level of the power amplifier in the companion audio processor. It acts as a low-pass (2,000 Hz) amplifier. Its output is then

rectified and fed into another μA702A stage, which acts as a level detector.

When the rectified input level to this stage exceeds a certain value, an output is created that is used to sharply reduce the gain of the audio power amplifier. Thus, only one audio channel will be amplified as long as it produces a sufficiently strong signal to retain control of the system at However, such a refinement would add little to the value of the unit for amateur DX receiving purposes, where the main objective is to select the receiver which, at any time, provides at least a minimum readable signal.

circuit description

The details of the circuit are shown in fig. 4 for one audio processor. The first

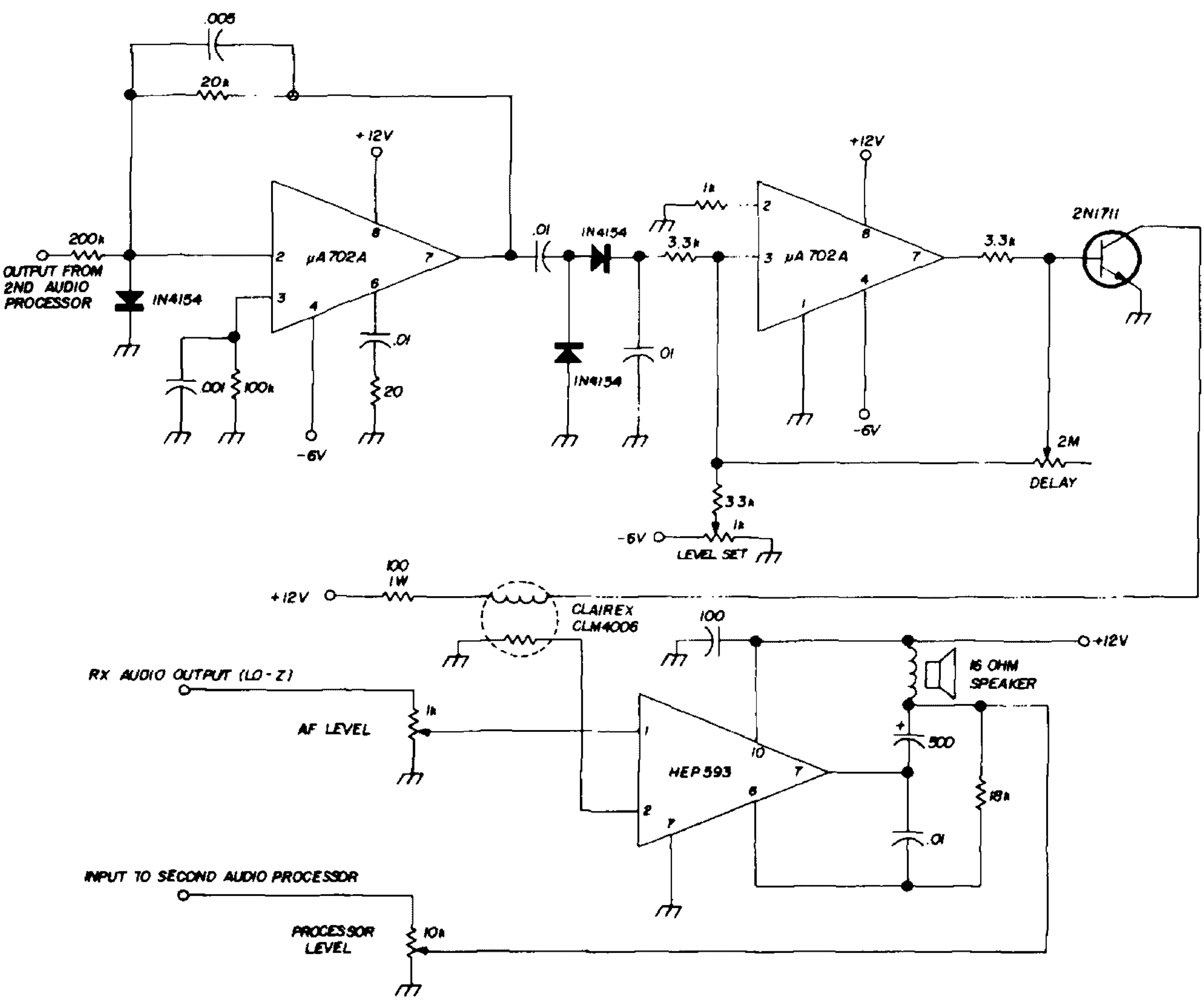
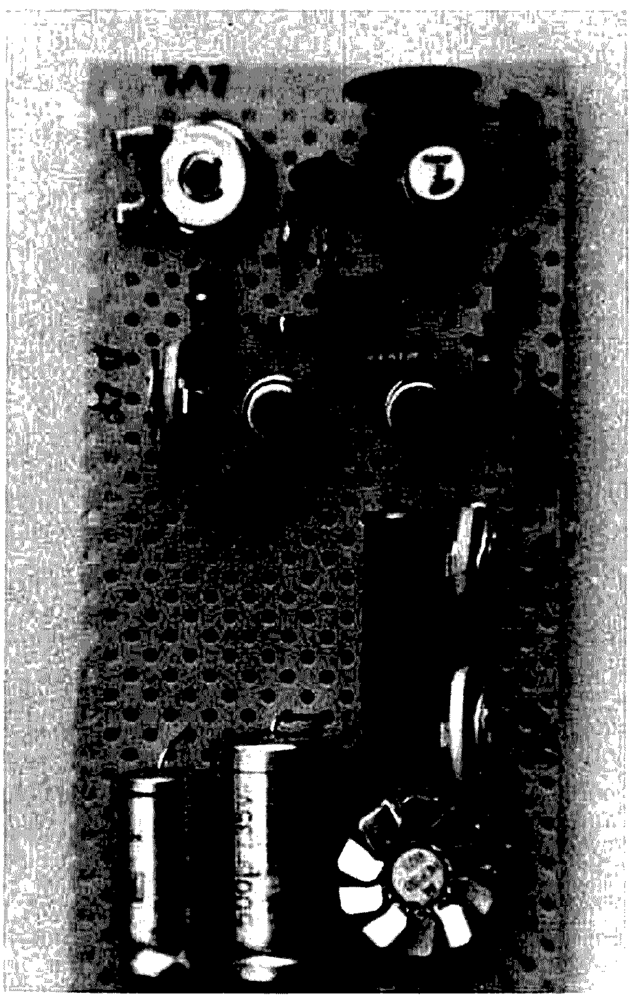


fig. 4. Complete schematic of one audio processor. Two Fairchild μA702As and one Motorola HEP593 are used.

the minimum level at which it has been set. If the signal falls below this minimum level, no output will control the system until a signal exceeds the minimum level. Only manual control of the system level is used. It is possible to provide automatic sampling of receiver levels to allow the stronger signal to set the system level. μA702A stage frequency roll-off was designed to hinder response to noise signals. The μA702A level-detector stage provides a positive output whenever the input level on terminal 3 exceeds that set by the *level-set* potentiometer. The positive output operates the 2N1711 driver transistor, which in turn switches the lamp on

and off in the Clairex 4006 photocell module. The resistive element of the photocell module controls the HEP593 amplifier gain. The audio amplifier is not completely cut off by the photocell, but its gain is reduced by about 40-50 dB. This, plus the thermal action of the photocell lamp, provides a smooth switching action instead of a thump each



Typical parts layout for an experimental audio processor. The HEP593 is shown at top (with cooling fins). The two μA702A's are below. PC-board pots are used for control functions shown in fig. 4.

time one or the other audio channel is activated. A delay loop is also included in the second μA702A stage and is controlled by a 2-megohm pot. This loop, plus the rectifier circuit between the μA702A stages, prevents loss of control of the audio-amplifier stage during short-term periods of low input to the first μA702A; e.g., during brief speech pauses.

construction

The construction of the processor units is identical. Perforated-board stock may be used as shown in the photo. The pots are PC-board types, but they should be installed as panel controls. The two processor units can be packaged into a complete assembly to suit individual requirements. Separate speakers were used for each audio channel instead of using isolating stages to drive a combining audio amplifier. The cost of the components, compared to that of an additional simple speaker, didn't justify the added complexity.

The power for the two units is −6 volts at 50 mA and +12 volts at 400 mA, which can be supplied by a well-filtered transistor power supply.

adjustment

When using the audio combiner, each receiver should first be adjusted to provide the same output level for a steady input signal. This can be done with the receiver controls and the *af-level* control on each processor unit. The *processor-level* control should be set to the ground end. The *processor-level* and *level-set* controls on each processor should then be advanced equally so that one receiver just controls the system within the smallest difference in control settings between processor units. That is, a small adjustment of the *level-set* pot on either processor should cause one of the audio channels to be activitated. The *delay* control should be set under actual receiving conditions so that excessive switching back and forth between audio channels, when no marked difference in output level between channels is noted, does not occur.

reference

1. Keith Henney, "Radio Engineering Handbook," McGraw-Hill, New York, 1959, pp. 19-118; 24-19; 24-25 to 24-31.

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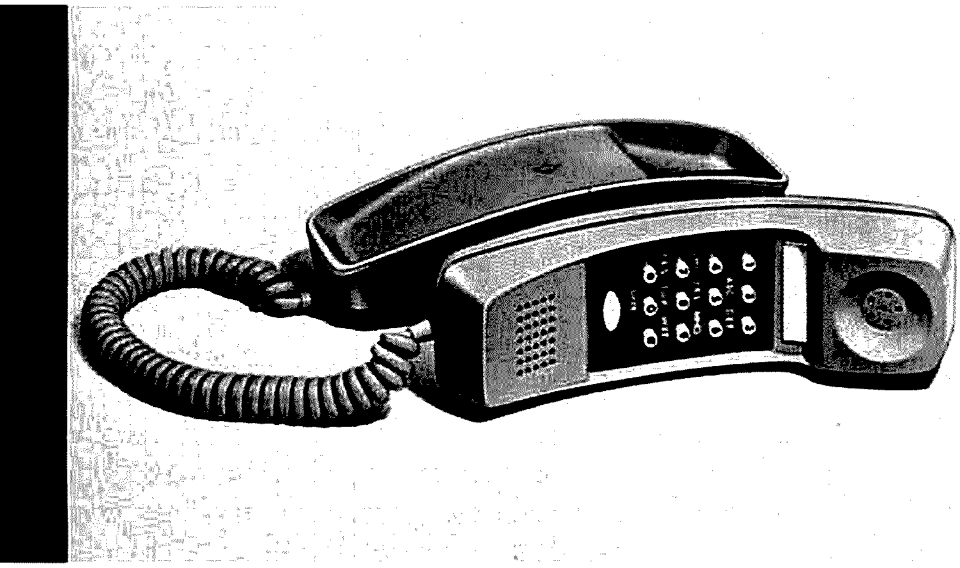


push-to-talk

for a styleline telephone

This simple solid-state circuit provides complete push-to-talk operation with the built-in recall switch included on Trimline and Styleline telephones

John C. Tirrell, W1DRP, 164 Cypress Lane, Nashua, New Hampshire 03060

Since many of the two-meter fm repeaters within range of my base station are either controlled by Touch-Tone, or about to be, I felt that I had to become better equipped. All I really wanted was a Touch-Tone pad, but I found that just the pad (keys and tone generator) were hard to get, although obtaining a complete Touch-Tone telephone was relatively easy.

Why not get the entire phone? Perhaps I could extract the needed Touch-Tone pad. I was confident that I could find a use for the leftovers. The trim new Styleline handsets manufactured by Automatic Electric have a Touch-Tone pad built into the handset along with a recall switch. The recall switch can be depressed momentarily to disconnect your call.

I figured I could use the built-in recall switch for push-to-talk. I wouldn't even have to remove the Touch-Tone pad; I could use the entire handset as a microphone. Little did I realize that it wasn't as simple as all that! One of the members of my radio club pointed out that the

This circuit can also be used with a Trimline telephone handset. editor.

built-in recall switch should be used for push-to-listen, *not* push-to-talk, since this button is in series with the cradle switch and completely disconnects the handset from the line.

One of the club members had taken his phone apart and mounted a *normally-open* push-button switch in the handset. He used the dial-light wires already in the cord. But what a job; the handset was so jam-packed that he could hardly get it back together again.

Dismayed but not undaunted, I began to think about solving this problem. Certainly I could activate a relay when I took the handset off the hook. This would turn the transmitter on; when I wanted to listen I could either push the recall button down and hold it, or hang up. The circuit for this arrangement is shown in fig. 1.

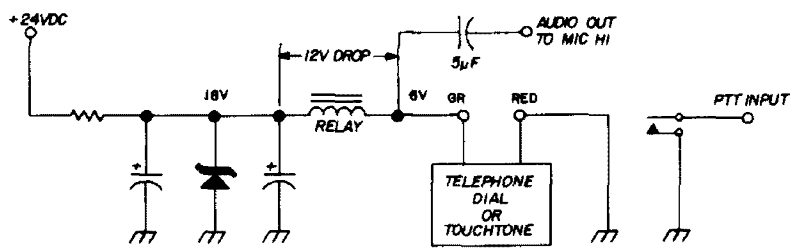


fig. 1. Automatic transmit control for a telephone set.* With this circuit the relay is energized when the handset is removed from the cradle. The relay contacts may be used to turn on the transmitter or activate a vox circuit.

*Styleline telephones are available from Junction Distributors, 164 Cypress Lane, Nashua, New Hampshire, 03060. Styleline Touch-Tone handsets are $39.20; Styleline base, desk type, $13.25; Styleline base, wall type, $12.45; Touch-Tone pad only, $27.20; enclosure for Touch-Tone pad, $6.80. Add $1.00 for orders west of the Mississippi River. Styleline telephones are available in beige, ivory, blue, ebony, green, yellow, avocado, pink and white; the Touch-Tone enclosure is available in either beige or black. Junction Distributors will provide a complete list of telephone equipment upon receipt of a self-addressed, stamped envelope.

When the new Styleline phone arrived a quick examination convinced me that holding down the recall switch to listen — not an easy button to operate — was not the way to go. I didn't even think I wanted to operate it as a push-to-talk *switch, trying to hold it down during* one of my long-winded transmissions.

I recalled the ordinary dial telephone: When the handset is picked up, a switch connects the telephone to the line. However, when a number is dialed, the line is momentarily broken the same number of times as the number dialed. What prevents line disconnection during dialing? "Delay," I thought to myself. The cradle switch is operated longer (typically much longer) than the dialing pulse mechanism.

I decided I would need two relays: one for control, the other for push-to-talk. The control relay, activated as soon as the handset is lifted from the cradle, would have a fast-operate slow-release characteristic so it would ignore the momentary operation of the recall switch. The push-to-talk relay would be controlled by the recall switch on the handset.

The control relay would provide two functions; when the handset was still on the cradle the relay would insure that the push-to-talk output could not be accidentally triggered, and it would make sure that the control circuit would always go to the receive mode when I hung up

the telephone.

In some installations the push-to-talk relay may not be needed. The push-to-talk relay in the transceiver could be driven directly from by the relay-driver transistor. However, be sure to put a diode across the relay coil to shunt transistor-destroying voltage spikes caused by the collapsing magnetic field when the coil drops out.

Since I planned to operate the recall switch as a momentary push-to-talk, momentary push-to-listen button, I needed a memory circuit; a J-K binary would be fine. The binary, actuated by the recall switch, would alternately change state, thereby providing driving logic for the sequential transmit/receive modes of the push-to-talk relay.

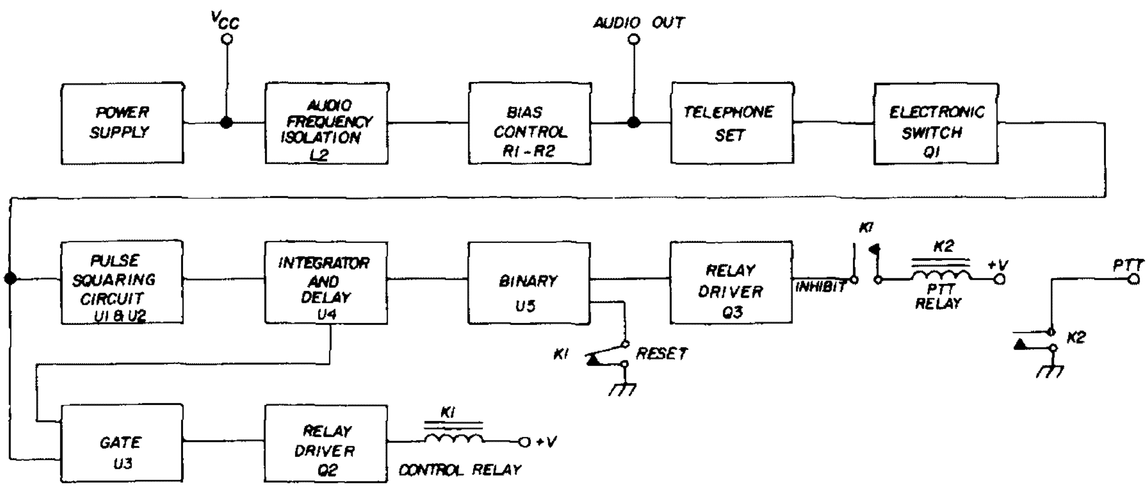


fig. 2. Block diagram of the push-to-talk control circuit. Complete schematic is shown in fig. 3.

power supply

Several different voltages were available because I was going to use the push-to-talk control assembly at my base station. Although 24 volts dc was available for the relays, instead of dropping this down to +5.0 volts for the integrated circuits, I decided to use a 6-Vac supply and build a small bridge rectifier circuit. This would give me 6 Vdc to operate the Touch-Tone and the carbon microphone in the handset.

If the control circuit is to be used in a mobile station, the +12 Vdc supply can be used with an appropriate voltage-dropping resistor in place of the bridge so that approximately +8 Vdc is available at the input of L1 in fig. 3.

All parts for the power supply came out of my junk box. Filtering is quite important since any hum on the power supply lines will show up as hum on the transmitted audio. The chokes I used were pulled out of old tv sets.

The voltage drop across both smoothing chokes must be considered when determining correct values for R1, R2

"As near as I can tell, it was frightened to death . . ."

and R13. The off-hook Styleline requires from 20 to 30 milliamperes of current. The lowest voltage at the input of R4 should not be less than 0.7 volts; otherwise Q1 will respond to audio inputs and cause erratic triggering. R13 is selected so that the power rating of the zener diode is not exceeded.

When I first put the circuit together I intended to put a capacitance delay circuit across relay K1. However, I soon discovered that the recall switch in the Styleline handset had considerable contact bounce. This generated multiple pulses each time it was operated with the result that it was impossible to determine the state of the J-K binary. I decided to concentrate on solving the contact bounce problem and worry about relay delay later.

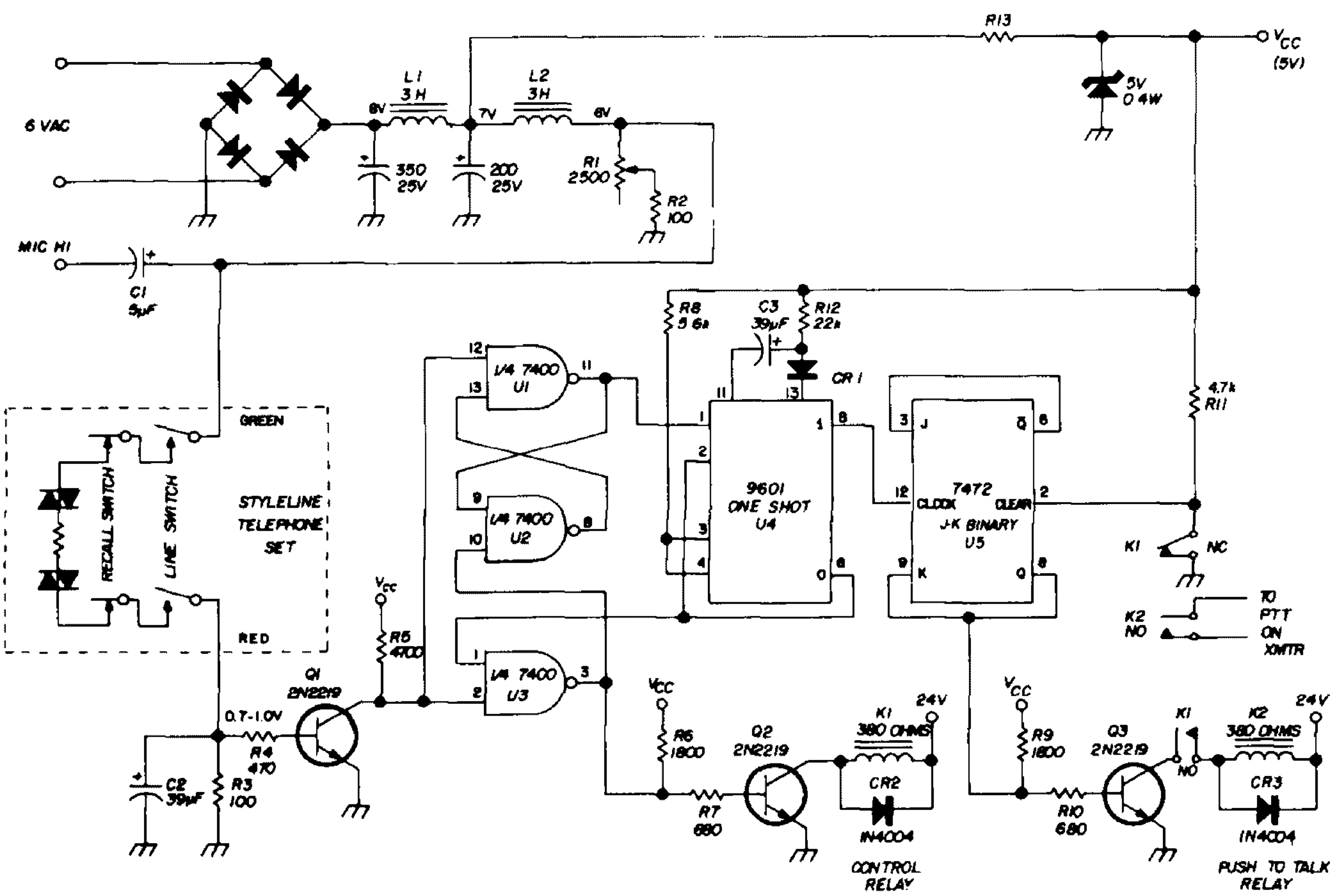


fig. 3. Push-to-talk control circuit for Styleline telephones is activated by the built-in recall switch. Integrated circuits U1, U2 and U3 are part of a Signetics 7400 quad two-input gate. U4 is a Fairchild 9601 monostable multivibrator. U5 is a Sprague 7472 J-K binary.

circuit

Inductor L2 provides additional filtering as well as providing a high impedance to audio signals. Capacitor C2 establishes a low impedance to audio signals across the biasing resistor R3. Both L2 and C2 are essential to the proper operation of this circuit.

I used 2N2219 transistors in the circuit because they were available. Any npn switching transistor that can withstand the relay driving voltage and provide adequate driving gain should work.

By adding a Fairchild 9601 integrated-circuit one-shot to the circuit and wiring the 0 output back to pin 2, a single pulse was produced at the output although many pulses appear at the input. The one-shot is triggered from a cross-coupled-gate pulse-squaring circuit; the one-shot output provides a clean clock trigger for the J-K flip-flop.

timing

The one-shot timing circuit (C3, R12 and CR1) is critical only to the extent that there is sufficient delay to insure that switch contact bounce has ceased. The delay time of the one-shot circuit can be determined from the following formula

$$T \approx 0.36\ RC(1 + \frac{0.7}{R})$$

where R is in kilohms, C is in picofarads and T is in nanoseconds.

Diode CR1 is used when capacitor C3 is an electrolytic. This diode prevents a reverse voltage across C3. The resistor for this circuit should be less than 30k, and all timing components must be as physically close to the circuit as possible to reduce noise pickup.

The one-shot provides an additional feature that soon became apparent. Rather than adding a delaying circuit across the control relay, the 0 output from the one-shot was used to maintain a holding level for the K1 relay driver during momentary operation of the recall switch. Although a short *off* pulse occurs (see fig. 4) it is much too short for K1 to respond.

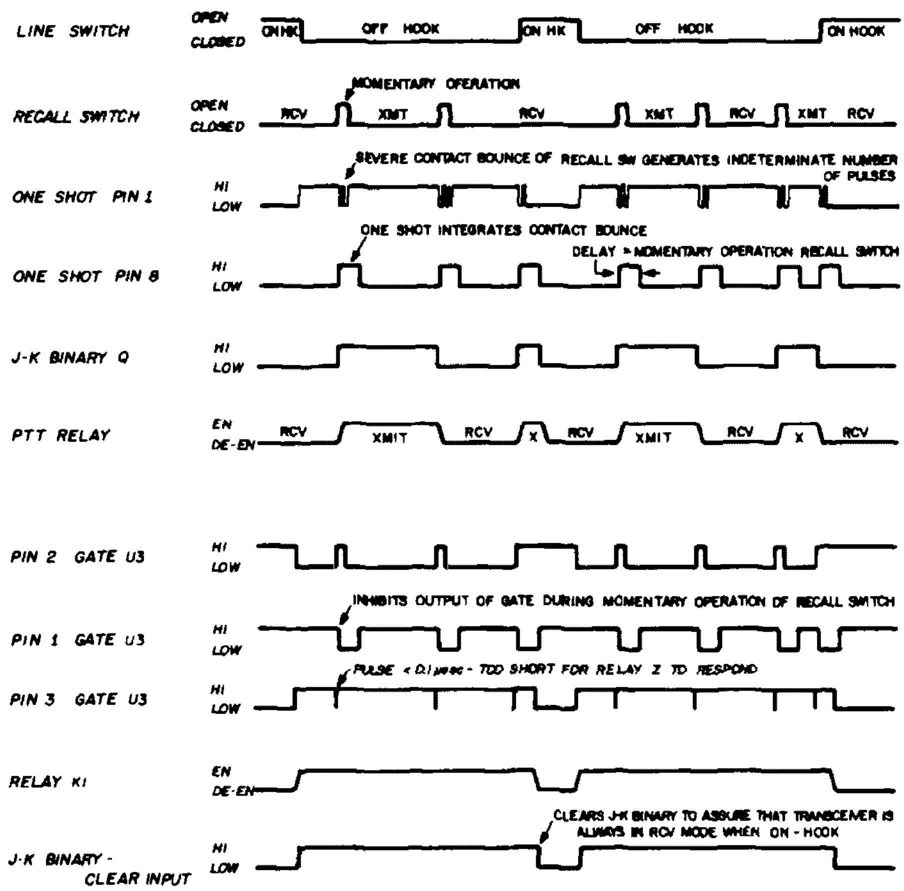


fig. 4. Timing diagram for the push-to-talk control circuit.

summary

I am using the push-to-talk *control* circuit in fig. 3 in my base station. When I hang up the Styleline handset the radio is switched to receive and stays in that mode until I pick up the handset and momentarily push the recall switch. It works very well, and provides a good deal of satisfaction when I show it off to those disbelievers who said it couldn't be done.

ham radio

frequency dividers for ssb generators

A simple frequency-dividing scheme for generating upper- and lower-sideband mixing frequencies for surplus ssb filters

Ken Stone, W7BZ, 641 Grant Avenue, Twin Falls, Idaho 83301

Some of the mechanical and crystal-lattice filters obtainable from surplus and other sources are not available with the necessary crystals for carrier generation in a ssb exciter. These crystals should be chosen at a frequency 20 dB down the skirt of the filter frequency response. They are fairly expensive if purchased new and surplus crystals are never quite on frequency. Here is a way to solve the problem simply and inexpensively.

The block diagram (in fig. 1) shows a method of sideband generation derived from the "Sideband Package" described by W6TEU.[1] It uses only one crystal to generate the necessary frequencies to produce a high-frequency sideband signal that does not shift frequency when changing sidebands. The 500-kHz carrier generator crystal is used to produce a ssb signal at 500 kHz; harmonics of the 500-kHz signal at 1 MHz and 2 MHz are used to heterodyne the 500-kHz signal to 1500 kHz. Upper and lower sidebands are selected by the choice of the 1- or 2-MHz injection frequency in the second balanced modulator.

Fig. 2 shows that the same results can be obtained by using a high-frequency crystal and a system of dividers to get the proper injection frequencies. This circuit has the advantage that a lower-cost more readily available high-frequency crystal can be used; this crystal can be "rubbered" in frequency so that it can be placed on the frequency corresponding to the 20-dB point on the filter curve.

The schematic diagram in fig. 3 shows a practical circuit I used in my solid-state ssb exciter. The 1500-kHz ssb signal could be used with a 5- to 6-MHz vfo to produce a usable signal on the 40- and 80-meter bands. However, in my case the vfo frequency was chosen to produce a tunable ssb signal between 6 and 6.5

MHz. This signal is heterodyned with crystal-controlled oscillators to get to the desired amateur band.

A little arithmetic will show that similar schemes can be worked out for 1500-kHz amplifier. Small toroid coils should work as well.

The diodes in the balanced modulators were surplus computer diodes and careful matching was not found to be necessary.

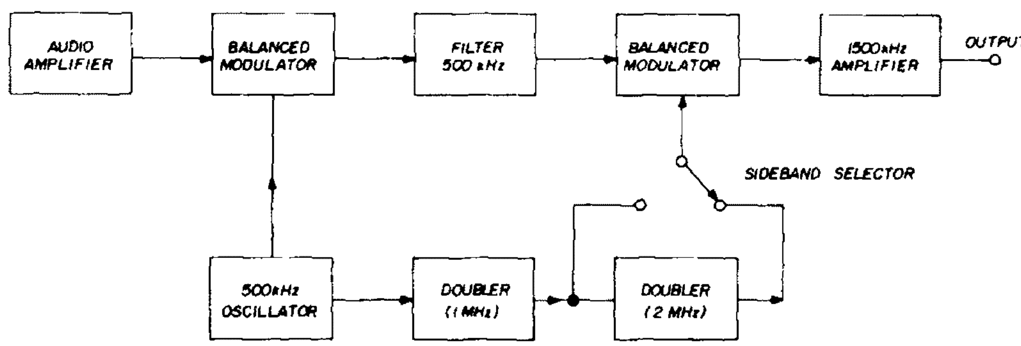


fig. 1. Block diagram of sideband generator derived from the Bigler's Sideband Package.

other filter and heterodyning frequencies by the proper choice of crystal and divider ratios. Integrated-circuit frequency dividers can provide almost any reasonable divider ratio.

construction

The sideband generator was easily built with parts at hand, and no special selection was found to be necessary. Any of the available JK flip-flops should work as well as the μL923s provided proper

The amplifier transistors are biased for class-A operation by proper choice of the resistors in the base circuit. A little instability was detected in the 2N2102 1500-kHz amplifier until the neutralizing circuit was added. This consists of a piece of wire wrapped around, but not connected to, the base of the transistor; the other end is connected to the free end of the plate tank circuit.

The crystal-oscillator circuit worked very well, and its output was sufficient to

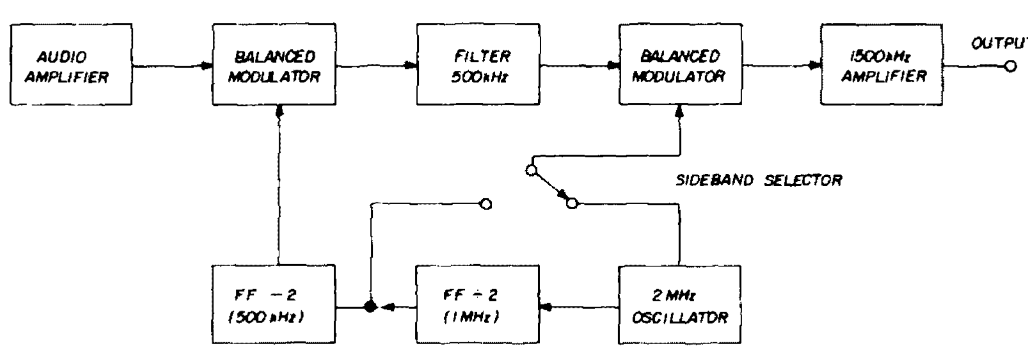


fig. 2. Modified sideband generator circuit uses frequency dividers for generation of upper- and lower-sideband carriers.

connections are made and the toggle frequency is not exceeded. The tuned circuits shown at the outputs of the flip flops were added to filter harmonics of the divider frequencies. These tuned circuits were salvaged from Japanese transistor radios as were the ones in the

drive the first divider and the second balanced modulator without any shaping or amplification.

Other crystals with less activity may require higher voltage on the oscillator to produce oscillation. In this case it may be necessary to limit the drive to the first

flip-flop to prevent damaging it.

You may wonder about the particular filter shown on fig. 3. While a 4-kHz bandwidth would not be very desirable in

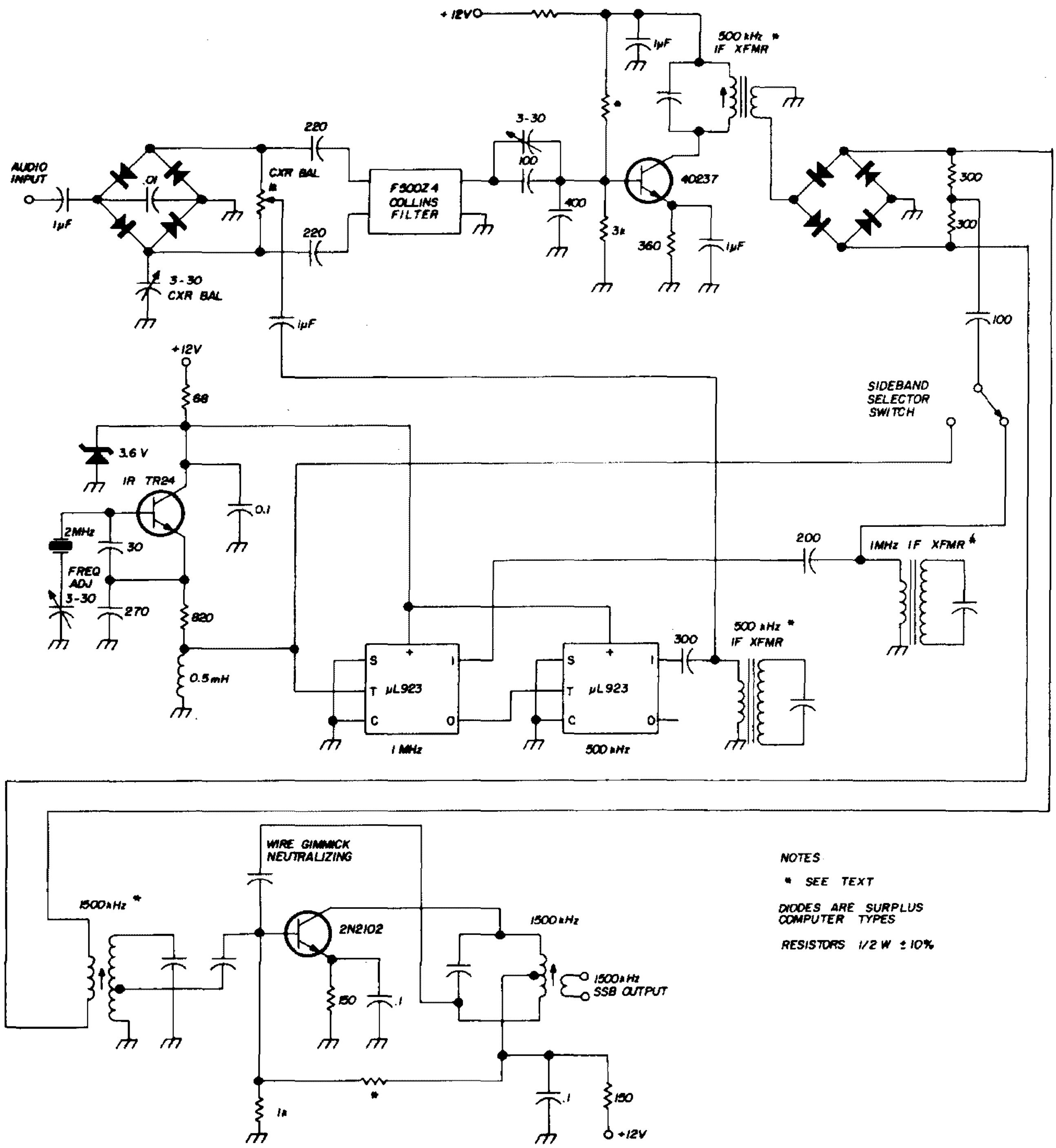


fig. 3. Schematic of the complete ssb package using IC frequency dividers. For components marked with an asterisk, see text. Diodes are surplus computer types.

a receiver or transceiver, it presents no problem in a simple ssb transmitter because the bandwidth is easily restricted in the audio system. This permits the use of mechanical filters of any bandwidth greater than 2.1 kHz as long as the filters have a steep skirt and desired out-of-band characteristics.

I hope that the ideas presented here will help other amateurs put some of the surplus filters to work in a rewarding homebrew project.

reference

1. A. A. Bigler, W6TEU, "A Sideband Package," *QST*, June, 1958, page 24.

ham radio

general-coverage receiver frequency calibrator

This circuit provides easily distinguishable frequency markers to above 30 MHz

G. Tillotson, W5UQS, 436 Grace Drive, Richardson, Texas 75080

For a number of years my station receiver was a ham-bands-only design. When I developed an interest in world-wide shortwave listening I found the amateur communications receiver too restrictive frequency-wise, so I purchased a medium-priced communications receiver.

The first problem was dial calibration; there were marks for each MHz point, but how did you locate an intermediate point? The usual 100-kHz calibrator was no help. When the distance between the 9- and 10-MHz calibration points is less than 1/8-inch, how do you determine which is the 9.7-MHz marker?

The circuit described here was designed to locate the desired marker. By starting at 1.0 MHz for the main dial marks, and switching to 500 or 100 kHz, it is easy to locate *any* 100-kHz point on the main tuning dial. In addition, this circuit provides strong harmonics well above 30 MHz.

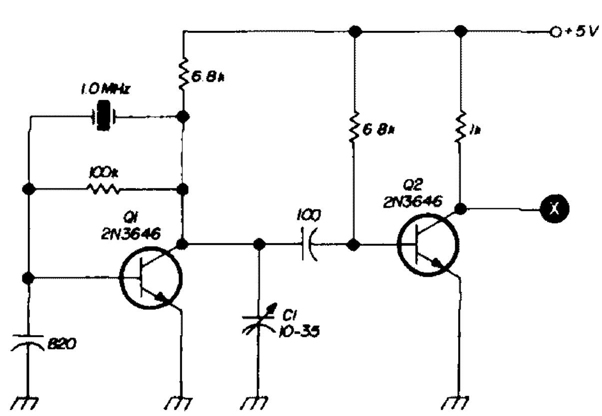
oscillator

The integrated circuit used in the crystal oscillator circuit in fig. 1 is a hex inverter; three inverter stages are used.

table 1. Operation of the integrated-circuit crystal oscillator.

crystal frequency		no padder	35 pF padder
1.00 MHz	32 pF	—	1.000 MHz
7.68 MHz	32 pF	7678.205 kHz	7.680 MHz
10.00 MHz	32 pF	9994.308 kHz	10.000 MHz
10.00 MHz	series	9997.428 kHz	10.000 MHz
15.36 MHz	32 pF	15339.505 kHz	15.360 MHz

The circuit will work with crystals from about 700 kHz to above 15 MHz as indicated in table 1. Variable capacitor C1 provides a means of adjusting the oscillator to precisely the desired frequency.



Alternate transistor crystal oscillator circuit.

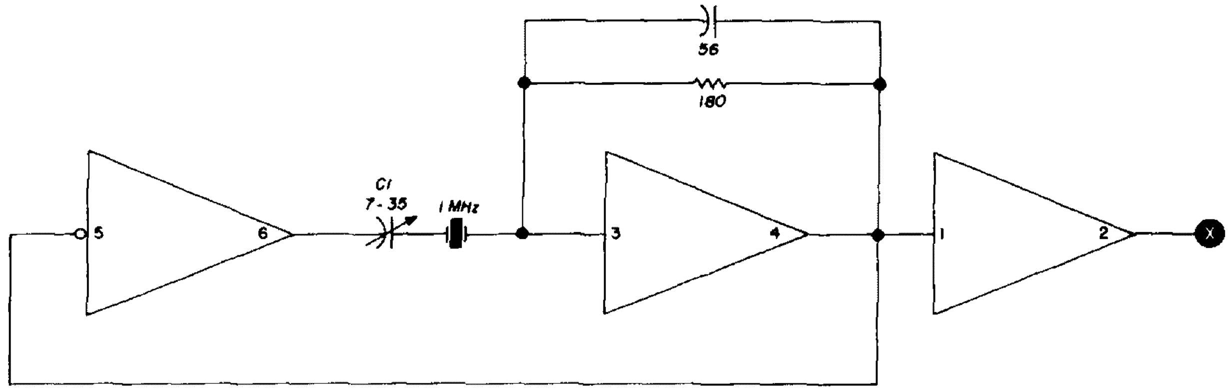


fig. 1. Integrated-circuit crystal oscillator circuit uses hex inverter, Motorola MC3008 or TI SN74H04. Alternate transistor oscillator is shown on page 28.

Since this is a fundamental crystal oscillator circuit overtone crystals oscillate at their fundamental. The output waveform is rich in harmonics. An alternate transistor oscillator is shown on page 28.

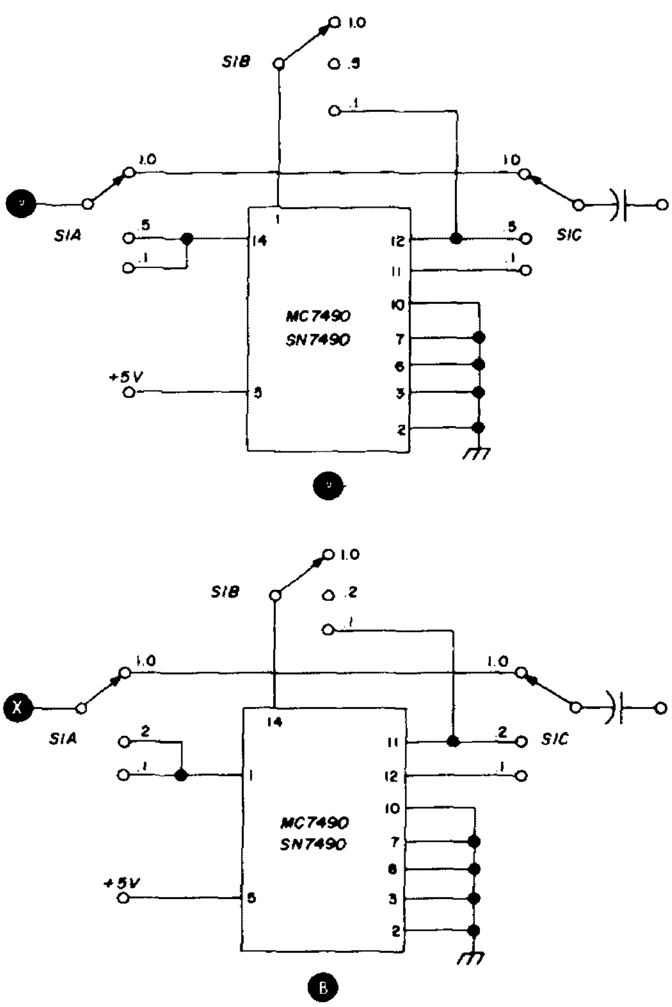


fig. 2. Frequency divider uses TTL integrated circuits. Arrangement in (A) provides output at 1.0, 0.5 and 0.1 MHz, while (B) provides output at 1.0, 0.2 and 0.1 MHz.

The two TTL integrated circuits used in the frequency divider are very fast. The circuit in **fig. 2A** is a divide-by-two section followed by a divide-by-five section.* With the connections shown in **fig. 2** two sets of frequency markers are avaiable, 1000, 500 and 100 kHz (**fig. 1A**) and 1000, 200 and 100 kHz (**fig. 1B**)

construction

The receiver calibrator is built into a small mini-box with switches S1 and S2 (power) located on one end. I used a miniature 3-pole, 3-position lever switch for S1, but a rotary switch could be used. Power is provided by the low-voltage supply in the solid-state general-coverage receiver. A portable supply could be made from three 1½-volt dry cells.

operation

With this circuit I had no problem with the oscillator starting or divider operation with the crystals I tried. The alternate transistor oscillator circuit did require capacitor substitution to get close enough for the trimmer capacitor to set a zero beat with WWV. For highest accuracy the unit should be calibrated against WWV with the divider set to the 100-kHz position.

ham radio

*Other divide ratios can be obtained by internal or external gating. For more information see Texas Instruments Bulletin CA-102, "Counters and Shift Registers."

integrated-circuit single-sideband speech processor

This high performance speech processing circuit provides more than 4 dB intelligibility threshold improvement with no distortion

Gene A. Nurkka, VK9GN, Box 73, Ukarumpa, E. H. D., Territory of New Guinea

Speech processing is one of the most popular topics for discussion among groups of amateur radio operators. When DXers gather at club meetings, sooner or later the talk turns to speech clippers, speech compressors, rf clippers and alc circuits. This is only natural. When the active DXer has put up the best antenna he can afford, as high as possible, and is using maximum legal power, the only way he can further improve his signal is through some method of speech processing.

Many excellent articles have been written on this subject. I highly recommend that you carefully review the articles listed in the bibliography. W6JES, W1DTY and K1YZW set forth the basic facts of speech processing and indicated that rf clipping is the best system for ssb communications work.[1,2] K6KA clearly shows that background noise must be completely absent when using any type of speech processing.[3] K1YZW shows why simple speech clipping should not be used with ssb equipment.[4]

time constants

In an audio-frequency speech compressor the ideal attack time should be less than 1 millisecond, and the release time should be faster than 10 milliseconds. However, the problem with such short time constants has been distortion which is caused by phase shift within the control loop. With the very complex internal circuitry of the IC used in the speech compressor in fig. 1 no phase-shift problems are encountered when using even shorter time constants. As long as the release time-constant is longer than several cycles of the speech waveform, there is very little distortion.

performance

The *only* way to evaluate accurately the actual improvement offered by a speech processor is to measure the Intelligibility Threshold Improvement (ITI).

To do this, a test path is set up where the unprocessed ssb signal is just barely understandable in the presence of white noise. The speech processor is switched in and a calibrated attenuator inserted in the line to the communications receiver. Attenuation is increased until the processed ssb signal is again just barely understandable; the value of attenuation is the ITI in dB.

While running ITI tests between my station and the club station, VK9UC, I found that the actual ITI measurement was very subjective: Under the same operating conditions different operators come up with different values for ITI.

The IC speech compressor in fig. 1 measured approximately 4-dB ITI with 18-dB peak limiting. With this circuit peak limiting can be as great as 28 to 30 dB before objectionable distortion occurs. With ordinary long-time-constant speech compressors ITI is only about 1 dB. After running these tests I feel that many authors are overly optimistic when quoting up to 10-dB ITI with a 20-dB peak-limiting rf clipper.

My reasons for selecting this type of speech processor are ease of installation and adjustment, satisfactory ITI without distortion and low cost. On all of these points the compressor is better than adding an rf clipper to your transmitting setup. Many of the rf clippers I have heard on the air produce noticeably more distortion than this circuit. This may be due to the adjustment and installation problems encountered when adding an rf clipping circuit to an existing ssb transmitter.

circuit

The National LM170, LM270 and LM370 are a family of IC amplifiers with built-in agc systems.* For amateur use the LM370 is completely adequate; the LM170 and LM270 are more expensive devices that have tighter operating specifications and extended temperature ranges.

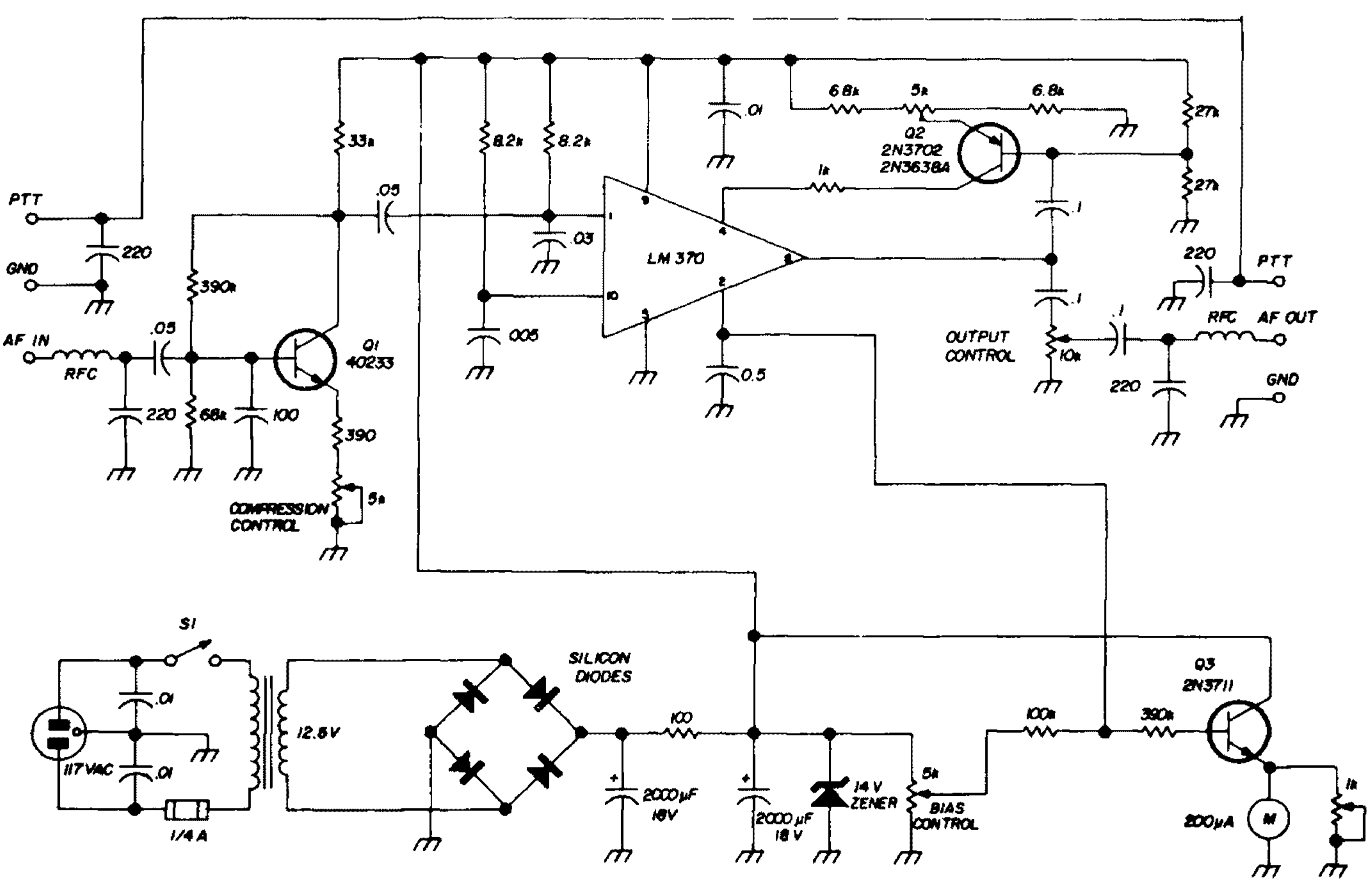


fig. 1. Schematic diagram of the IC audio speech compressor.

The LM370 amplifier circuit, built into a TO-5 transistor package, contains 34 transistors and 20 resistors.

Transistor Q1 in fig. 1 is used in a low-noise, low-gain stage which amplifies the microphone input slightly to drive the LM370 IC into its compression range. The 5k *compression control* is a variable emitter resistor which controls the gain of the stage. This effectively varies the amount of compression; maximum resistance coincides with minimum compression.

The input impedance of this stage is sufficiently high for use with high-impedance microphones. I normally use a low-output 200-ohm dynamic microphone, and there is sufficient range to provide plenty of compression.

The input, output and power leads are filtered for rf as are the base of transistor Q1 and pins 1 and 10 of the LM370 integrated circuit. The 0.03-μF capacitor connected to pin 1 of the IC provides some roll-off for high audio frequencies.

The 0.5-μF capacitor from pin 2 to ground determines the attack and release times of the circuit. Attack time is approximately 0.5 millisecond, and release time is about 10 milliseconds. I found during compression that there was quite a bit of negative-peak distortion due to charging this capacitor to 1.5 volts (see fig. 2). When the voltage on pin 2 is less than +1.5 volts, there is no compression, so I biased the capacitor to about the +1.5-volt level. This completely cleared up the distortion as shown in fig. 3.

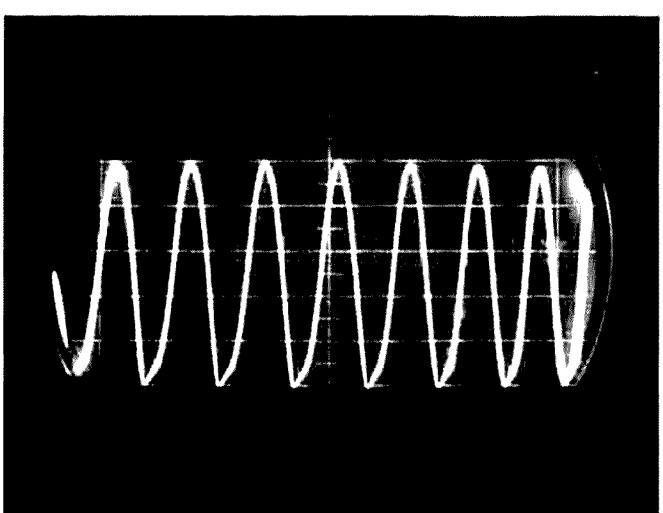


fig. 2. Negative-peak distortion with 18-dB compression (see fig. 3).

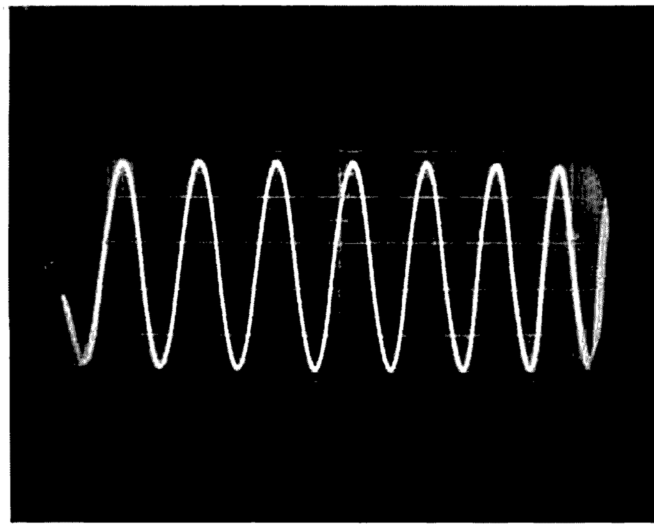


fig. 3. By biasing pin 2 of the LM370 IC at +1.5 volts, there is no negative-peak distortion at 18-dB compression.

Since the voltage on pin 2 is related to the compression level I added a dc amplifier (Q3) and a 200-microamp meter as a compression indicator. To set up the meter circuit, disconnect the 390k resistor from pin 2 and connect it to +1.5 volts dc. Adjust the 1000-ohm potentiometer until the meter reads some convenient level about two-thirds full scale. Reconnect the 390k resistor to pin 2.

Now, with the compressor on, but with no input signal, adjust the 5k *bias control* until the meter reads to the same two-thirds point or slightly less. This reading will drift slightly with ambient temperature changes. This is all right as long as the meter does not go over the two-thirds point (pin 2 not over +1.5 volts).

In normal operating the meter will flick somewhat higher with compression. The higher the meter reading, the greater the compression; with no movement of

*The LM170, LM270 and LM370 agc/squelch amplifier integrated circuits are manufactured by National Semiconductor Corporation, 2975 San Ysidro Way, Santa Clara, California 95051. The LM370 is priced at $4.50 in small quantities.

the needle, there is no compression although the unit still operates as a preamplifier. Some high-output microphones will produce occasional voice peaks which will be compressed even with the *compression control* turned fully down.

Transistor Q2, along with the internal circuitry of the IC, provides detection of the negative audio peaks. There is no current flow through Q2, and no compression, until the negative voice peaks exceed a certain level.

The base of Q2 is biased to one-half the supply voltage. The emitter voltage is set with the 5k potentiometer to provide an output from zero to several hundred millivolts. If this pot is set too near the zero-output point a severe transient pulse appears at the output of each syllable. Other than this the setting, this potentiometer is not critical. With the wiper of the 5k pot set at its lower end you will obtain maximum output and a compression range of about 28 dB.

If you have an oscilloscope and an audio signal generator you can experimentally check various settings of this 5k emitter potentiometer. I have found that there is one point where you obtain

I included a simple ac power supply rather than using batteries. When I first built the unit I used batteries, but I often forgot to turn the compressor off when I closed down the station, and in only a few days the batteries would be dead. If you're only going to use the compressor for amateur communications (I also use mine when tape recording speech) the +14 volts may be taken from your transmitter power supply. The zener diode is not really necessary and may be left out, although the supply voltage will increase slightly.

My compressor is built into a small metal cabinet as shown in the photographs. Two printed-circuit boards, shown in **fig. 4**, are used. One circuit board contains the IC-compressor circuit; the other holds the final filter capacitor and metering circuit.

operation

When my wife and children, with their high-pitched voices with many peaks, speak into the microphone, the compressor is worth its weight in gold. With a normal male voice the compressor provides at least 4 dB improvement. Al-

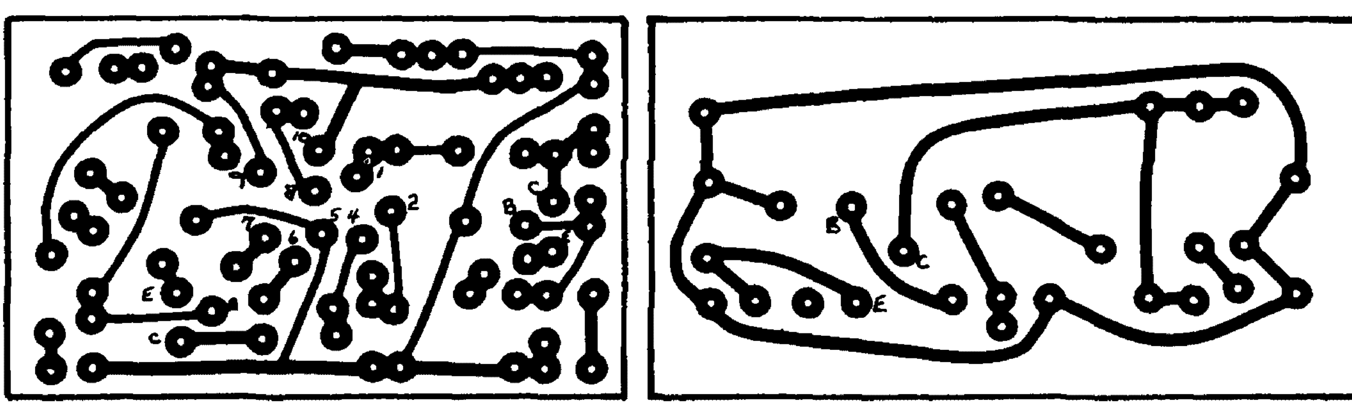


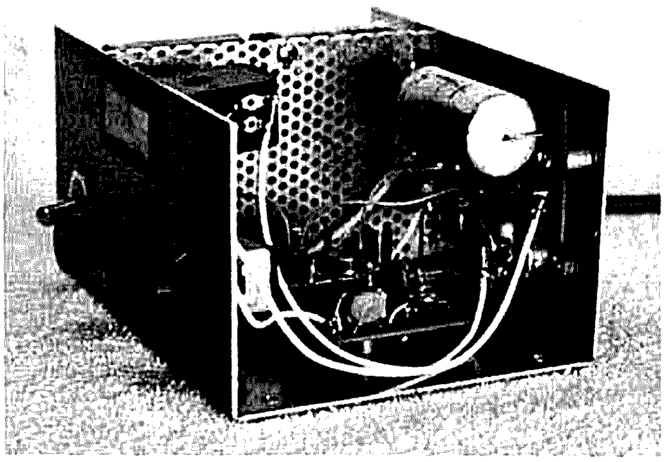
fig. 4. Full-size printed-circuit boards for the speech processor. Board on left contains IC and associated components; board on right holds final filter capacitor and meter circuit.

maximum overall compression range before distortion occurs, and this varies from one LM370 to another. The improvement in compression range in only a matter of a few dB so it may not be worth the effort.

Since the circuit requires about 13 mA

though this is substantial it is *not* going to make you top dog in the DX pileups. Although your peak power remains the same, a 4-dB improvement in peak to average power ratio is approximately equivalent to increasing your transmitter power from 100 to 250 watts.

If your transmitter power supply sags with this higher output power the speech compressor will not provide much on-the-air improvement. Also, if your transmitter uses tv sweep tubes in the final



Construction of the speech processor. Power supply is located behind the shield.

amplifier you will probably be replacing them more often because the higher average power will result in greater heat dissipation and shorter tube life. This is where the *compression control* and meter come in handy; when signals are good I usually run the compressor so the meter only occasionally flicks upward, indicating occasional compression of voice peaks. This gives maximum tube life and prevents overdrive and splatter.

When my wife and children are talking I use more compression, and they get through solidly. When signal conditions are bad, or rare DX appears, I use maximum compression to provide maximum talk power.

references

1. H. G. Collins, W6JES, "Ordinary and Processed Speech in SSB Application," *QST*, January, 1969, page 17.
2. J. R. Fisk, W1DTY, "Speech Processing in Amateur Equipment," *ham radio*, June, 1968, page 60.
3. E. H. Conklin, K6KA, "Audio Compressors and Clippers," *ham radio*, April, 1969, page 24.
4. W. Schreuer, K1YZW, "Speech Clipping in SSB Equipment," *ham radio*, February, 1971, page 22.

ham radio

electronic speed conversion for RTTY teleprinters

A reliable and accurate alternate for gearbox speed translators is this converter using ujt clocks and IC counters

Laurence H. Laitinen, WA6JYJ

Many RTTY operators are interested in using their teleprinters at speeds higher than 60 wpm for activities such as punching tape (e. g., operating a model 19 teleprinter* at 75 wpm to obtain hard copy while punching), MARS work, snooping around the commercial frequencies for press copy, or for possible multi-speed operation in the amateur bands should a current proposal before the FCC be adopted.[1]

Such operation requires at least two machines, one operating at 60 wpm and another for higher speeds. If more than one higher speed is of interest, more machinery will be needed. Most RTTY operators use the higher speeds only occasionally so the high-speed machines are idle most of the time. Even with

*This equipment and other RTTY gear is discussed in "First Steps in RTTY" by Chuck Schecter, W8UCG, which appeared in the June, 1971 issue of *QST*. editor

multiple-speed authorization, 60 wpm will probably be the predominant RTTY speed, since most machines currently in use operate in this range.

Owners of model 28 machines may use a gearbox to print at two or three different speeds – if their pocketbooks can stand it. However, such items are expensive and hard to come by through surplus sources.

An alternate approach, suggested by W6JVE, is to use electronic speed conversion. A few years ago this method would have been impossible. Today, ICs and other sophisticated solid-state devices make electronic speed conversion easy and relatively inexpensive. In the converter shown here, only 10 ICs and 7 transistors are used (excluding those in the power supply).

operation

The logic diagram of the speed converter is shown in **fig. 1.** Its input is connected to a terminal unit and the output to a model 28 printer operating at 100 speed (a model 15 may be used at 75 speed).

The speed converter receives slower-speed teleprinter signals and temporarily stores each character (one at a time) in a shift register. It then makes a parallel transfer to the output shift register and transmits the character to the model 28 printer at a 100-wpm rate. While the speed converter is transmitting the character to the model 28 printer, it is ready to receive another character. The receiving speed can be anything less than or equal to the converter output speed. Speed selection is by a switch that changes the timing resistance in the receiving section unijunction oscillator clock.

In addition to the speed-converted output, an additional output is provided that is equal to the receiving speed. This output is a regenerated replica of the input signal and can thus be used for bias-free retransmission.

Transistor-transistor logic is used throughout. The shift registers are five-bit registers (on one chip) plus a single type-D flip-flop to make the registers six bits long. The receiving section counter is composed of four type-D FFs. The original speed converter used a four-bit dtl counter when ttl prices were relatively high. Now, at the physical expense of another chip package, it's cheaper to make the counter from discrete FFs.

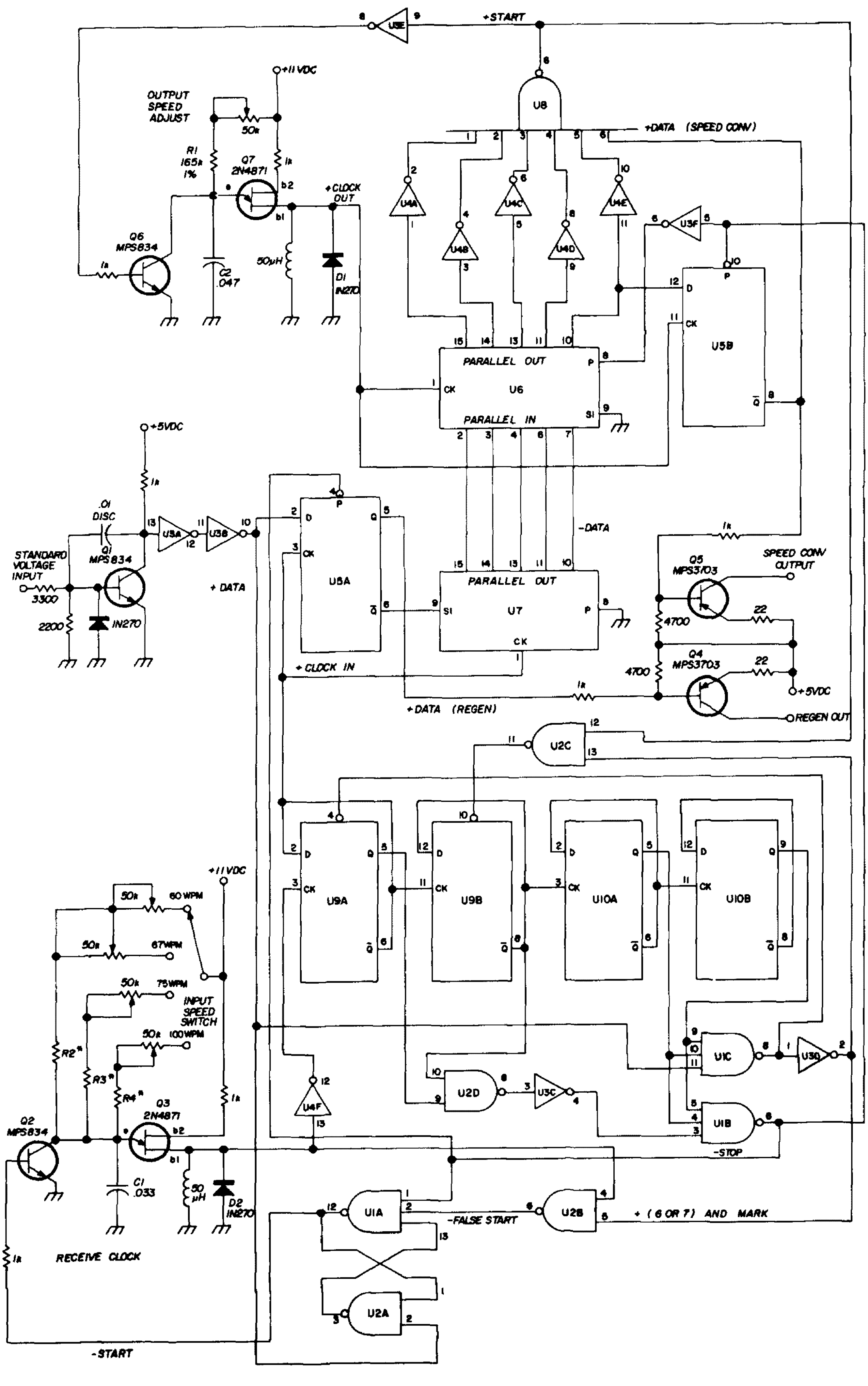
The input section consists of W6JVE's standard voltage interface.[2] (Alternate arrangements are shown in **figs. 2A** and **2B.**) Since the standard for mark and space is a negative and positive voltage level respectively, the output of Q1 is +mark and is thus labeled +data. The input impedance of TTL logic is fairly low, so Q1's output is buffered by inverters, two of which are necessary to preserve signal sense.

The input of the six-bit receiving shift register is half of a 7474 dual type-D flip-flop. This ff provides the regenerated output signal at the same speed as the incoming signal. Thus, its noninverted (Q) output is connected to the standard voltage interface output circuit. Although not shown, Q4's collector is returned to a negative voltage source through the standard voltage interface hub. An alternate output arrangement is shown is **fig. 2C.**

receive clock

The receiving shift register clock is controlled by a latch composed of cross-coupled 3-input and 2-input NAND gates (IC-1A, IC-2A). During a start pulse (space), +data goes low setting the latch and thus turning off Q2, the control transistor for the ujt clock, Q3. The timing capacitor, C1, now charges to the firing point of Q3 through one of the switch-selected timing resistor combinations. The period of oscillation is set to twice the Baud rate by the network RC time constant (see table 1).

The 50 μH choke in the base-1 lead of



+START
+11 VDC
OUTPUT SPEED ADJUST
50k
R1 165k 1%
Q7 2N4871
1k
+CLOCK OUT
Q6 MPS834
C2 .047
50μH
D1 IN270
+DATA (SPEED CONV)
U8
U4A
U4B
U4C
U4D
U4E
U3E
U3F
PARALLEL OUT
U6
PARALLEL IN
U5B
+5VDC
.01 DISC
Q1 MPS834
STANDARD VOLTAGE INPUT
3300
2200
IN270
U3A
U3B
+DATA
U5A
-DATA
U7
SPEED CONV OUTPUT
Q5 MPS3703
4700
22
Q4 MPS3703
+CLOCK IN
+DATA (REGEN)
REGEN OUT
U2C
U9A
U9B
U10A
U10B
60 WPM
67 WPM
75 WPM
100 WPM
INPUT SPEED SWITCH
R2*
R3*
R4*
Q2 MPS834
Q3 2N4871
C1 .033
50 μH
D2 IN270
U4F
U2D
U3C
U1C
U3D
U1B
-STOP
U1A
U2B
-FALSE START
+ (6 OR 7) AND MARK
U2A
RECEIVE CLOCK
-START

Q3 provides an output pulse width of approximately one microsecond. The 1N270 diode prevents the output from going negative after Q3 turns off.

counter

Since the receiving clock oscillation frequency is twice the Baud rate, the first pulse out of Q3 occurs halfway into the start pulse (11 ms for 60-speed reception). This pulse is applied to the counter (U9, U10) through an inverter so that the counter toggles to 0000 (previously 1111) on the falling edge of Q3's pulse if the received signal is still spacing. (Note that the 7474 ff is positive-edge triggered.) If the signal is not spacing at this time, the counter is inhibited from toggling, and -false start resets the latch. Assuming it is spacing, the first stage of the counter divides the output frequency of Q3 by two and provides clock pulses to the shift register. The inverted ($\overline{Q}$) output of the first counter stage goes from low (0) to high (1) half a bit-time into the start pulse and thus provides +clock-in to the shift register. After the clock pulse rises the data at the input of ff U5A shifts to its output. The input data is now free to change without affecting flip-flop output. The start pulse is now briefly stored by this flip-flop.

The operation above repeats for all five selecting pulses and the stop pulse of the incoming character. Each positive transition of the clock pulse shifts the data one stage to the right. Since the start pulse isn't needed in the speed-conversion process, the start pulse is allowed to shift out of the right end of the shift register so that the five selecting pulses are stored in the five-bit shift register, U7. Note that the input to U7 is from the inverted ($\overline{Q}$) output of ff U5A; thus the parallel

fig. 1. Speed converter logic diagram. The value of C2 is for output speed of 100 wpm (see table 1 and text). Except as noted all resistors are 0.25 watt, 10% tolerance. Capacitors C1 and C2 are 100-volt Mylar, 10%. Capacitor values shown may require up to 5000 pF of padding due to a wide variation in unijunction transistor characteristics and capacitor tolerance (see text).

table 1. Timing resistor values for various speed and Baud rates.

speed and Baud rates	input clock frequency	resistors*
60 speed 45.45	90.90 Hz	R2 = 174
67 speed 50.00	100.0 Hz	R2 = 174
75 speed 56.67	113.3 Hz	R3 = 147
100 speed 74.2	148.4 Hz	R4 = 110

Output clock frequency = Baud rate of desired output speed. Select C2 for desired output speed (e.g., 100 wpm, C2≈0.047 μF; 75 wpm, C2≈0.062 μF).

*1% tolerance

output of U7 is -data. The stop pulse is stored in ff U5A, which provides the stop pulse for the regen output.

sequencing

Each clock pulse from the first stage of the counter increments the last three stages of the counter by one. (We'll ignore the first stage of the counter for counting purposes since the first stage is not in decoding but only for the shift-register clock pulses.) The output of U1B (-stop) resets the clock control latch when the counter reaches 6 (110 counted from right to left in the counter). This normally occurs during the stop-pulse interval. Q2 is turned on by the reset latch, and thus the receive clock is inhibited from producing further output.

If it were not for gate U2D and inverter U3C, U1B would see a count of 6 when the counter turns to 0000 at the beginning of the receiving cycle when the first clock pulse is produced. This is due to the nonzero propagation delay of the counter ffs. Thus U2D inhibits the brief glitch that results when the counters turn to 0000 by first inputing a low signal to U2D several nanoseconds before U9B reaches 0. Note that U9B reaches 0 while U10A and U10B are still 1, and thus a false count of 6 is produced. U2D prevents this false count from reaching the 6 decode gate, U1B. When the counter actually toggles to a count of 6 U2D allows this count to pass to the

decode gate, since the first ff is set to a 1 (with the signal line marking during the stop interval). If the signal line is not marking, the next clock pulse will toggle the first flip-flop to a 1.

Thus the receiving portion of the speed converter stops during the stop-pulse interval regardless of whether the

completes the race loop, since the ($\bar{Q}$) output of the counter's second stage is now low, and the count is seven thus causing the output of the six decode gate, -stop, to go high, which is where the race started.

The speed converter receiving section is now ready to receive another character,

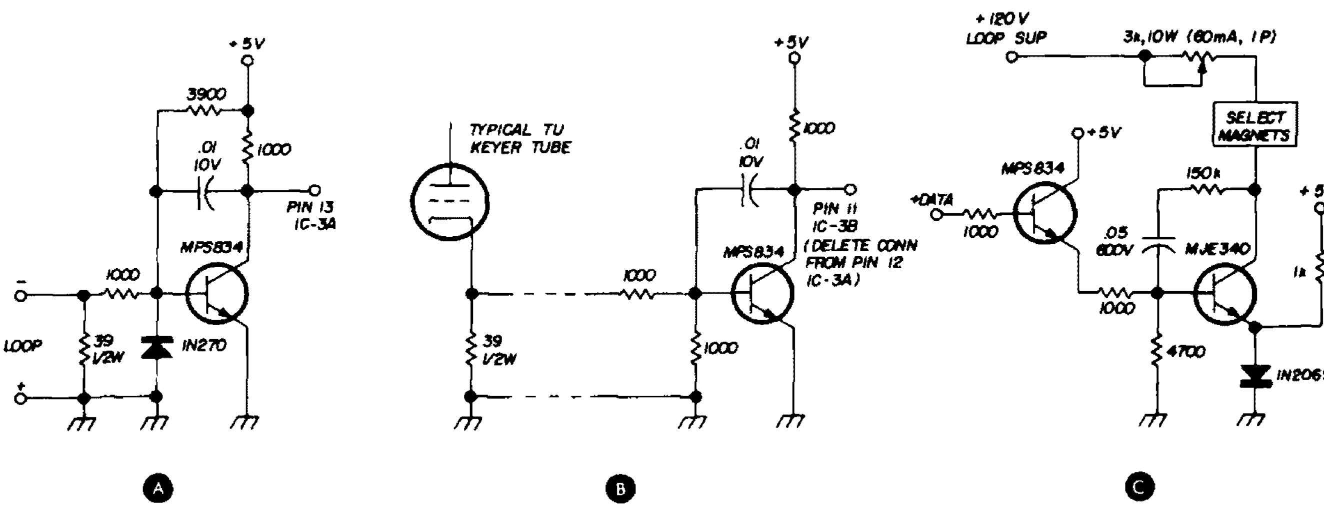


fig. 2. Alternate input circuits are shown in A and B; an alternate output circuit appears in C.

stop pulse is present. Such action is quite unlike a mechanical selector, which keeps operating if the stop pulse is missing, leading to synchronization problems.

We now have received the character, and the next step is to transfer it to the output register and reset the counter for the next receiving cycle.

At this point assume that a normal stop pulse is present. U1C detects +mark (from the +data line) AND +6 causing its output to go low, which presets the first stage of the counter to a logical 1.

The second stage of the counter is preset by a race pulse initiated by the -stop signal as follows: The low -stop signal presets the last stage of the output shift register (U5B). The ($\bar{Q}$) output of U5B goes low, which causes the output of U8, an 8-input (only 6 used) NAND gate, to go high. This signal, +start, is ANDED together with +6 AND mark from U1C and U3D to preset the second stage of the counter to a logical 1. This

since the -stop signal has been removed from the latch, and the received character stored by the input register has been loaded into the output register for transmission.

output circuit

The method used to control the output-section clock is to detect space signals in the register and keep the clock on until all spaces have been shifted out. Even if the received character is a *Ltrs* function, one space will be present because the ($\bar{Q}$) output of U5B is low as a result of the race pulse. This low output provides the start pulse and also turns on the unijunction oscillator clock by causing a high on the +start line. In the case of a *Ltrs* function, only one clock pulse is generated. The pulse occurs approximately 13 milliseconds (for 100 speed) after U5B has been preset by the race pulse. The clock pulse shifts the data one stage to the right and shifts in a -mark at the left

end (serial input is grounded, thus -mark).

When the start pulse has been shifted out the right end all the inputs to U8 are high, which causes +start to go low, and the clock turns off. If the character K had been received, for example, a space pulse would have been loaded into the first and the last stage of the output shift register. The space pulses move one stage to the right after each clock pulse until the last pulse is shifted out. At this time the clock is turned off because all the inputs to U8 are high.

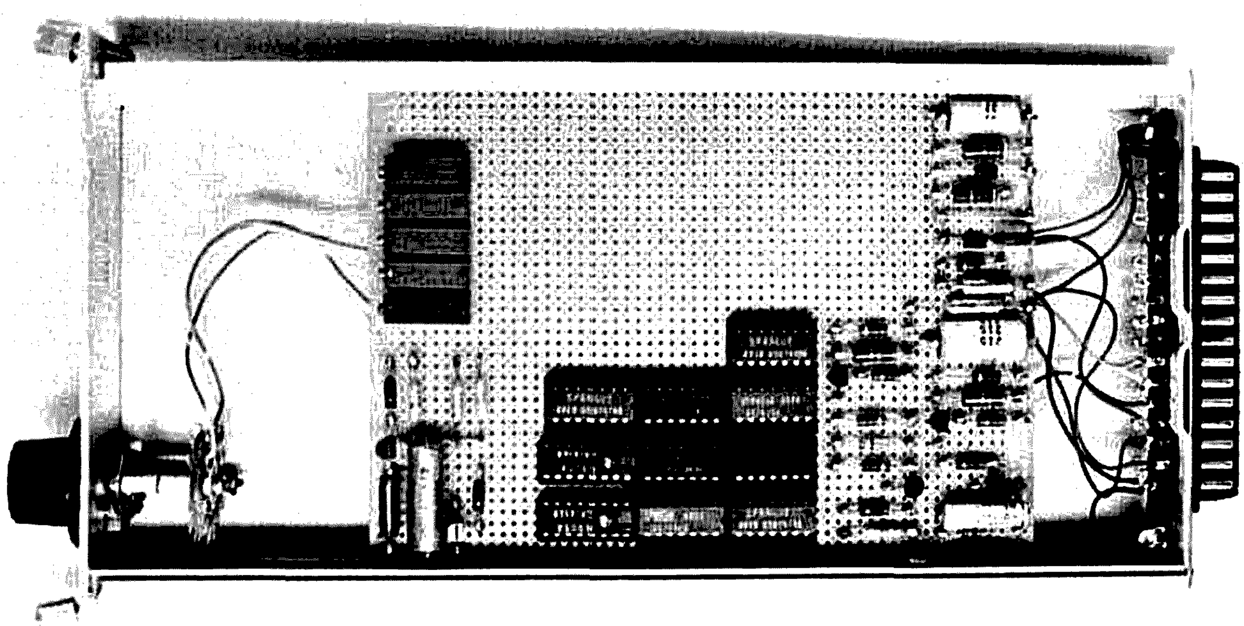
The output cycle is now complete and ready to receive another character from the input section.

construction

The ICs, transistors, and other discrete components were mounted on type 169P59/047 Vectorboard. Sockets can be used to mount the ICs but aren't necessary. The pins of the ICs can be pushed through the holes in the board and the interconnection wires soldered directly to them. Use Vector T-42-1 Micro-Klips to mount the discrete components.

All npn transistors are Motorola MPS834 or equivalent. Diodes D1 and D2 must be germanium — 1N270s are preferred. Silicon diodes cannot be used because of their high voltage drop, which allows the unijunction oscillator output to go too far negative when it turns off. Q4 and Q5 are Motorola MPS3703 or equivalent pnp transistors.

Although the noise immunity of TTL logic is better than RTL logic, TTL generates a bit more noise. This is especially true of the shift-register chips. It's a good idea to solder 0.01-μF disc ceramic bypass capacitors between the V_{ee} and V_{cc} pins of the shift register chips to prevent their noise pulses from being distributed to the other chips.



Internal parts arrangement of the RTTY speed converter. I used sockets, but the ICs could have been mounted directly on the Vectorboard with the leads pushed through the holes in the board.

A suitable power supply is shown in fig. 3. Be sure to insulate the 2N4921 from the heat sink (chassis) with the mica washers supplied with it. Verify that the power supply is working correctly before connecting it to the logic.

Five-percent resistors were originally used in the timing networks of the unijunction oscillators. However, considerable improvement in long-term stability was found through the use of one-percent precision resistors. Stability on the order of one or two Hz over a period of several months has been obtained with precision resistors and ten-percent mylar capacitors. Unlike resistors, the tolerance of mylar capacitors doesn't appear to be an indicator of their stability.

initial operation

Before applying power check the speed converter for wiring errors. This doesn't take much time and can eliminate a lot of problems that might otherwise be hard to track down.

If a frequency counter is available, connect it to the base-1 lead of (Q7). Ground the base of Q6 and adjust the associated pot for a frequency of 74.2 Hz for 100-wpm operation (56.7 Hz for 75 wpm).

Due to a wide variation in unijunction transistor and capacitor tolerances, the value shown for C2 is minimum. It may be necessary to pad C2 with a few thousand pF to place the desired clock frequency in the center of the output pot's range. If a counter is not available, use a calibrated oscilloscope. Lacking this, temporarily remove the preset input to U9A (pin 4), ground the base of Q2 instead of Q6, place a mark signal on the input of the converter, then slowly turn the pot until the receiving machine prints *Ltrs* and then something other than *Ltrs* function. Place the pot midway between these two points. Remove the ground from Q6 if still present and replace the preset input lead to U9A if removed.

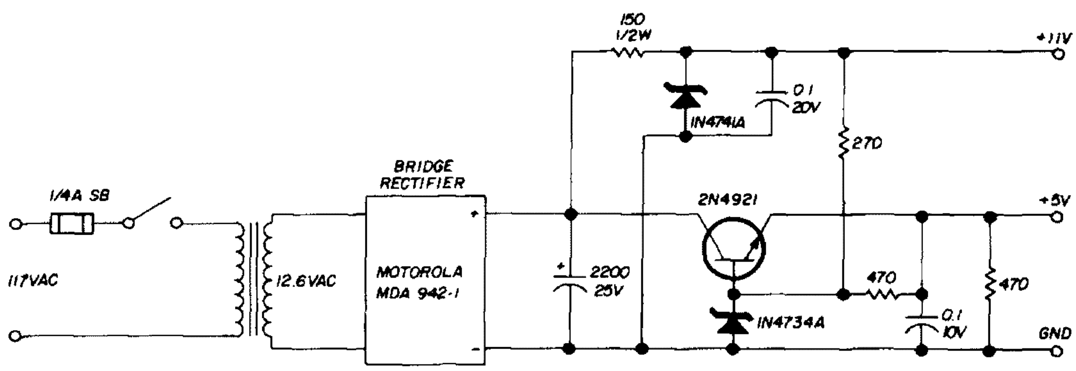


fig. 3. Suggested power supply. The 2N4921 transistor is bolted to the chassis (use mica insulating washers).

receive clock adjustment

The receiving clock frequency should equal twice the Baud rate of the selected receiving speed. Place the input-speed switch to the 60-wpm position, ground the base of Q2, and adjust the 60-wpm pot for 90.0 Hz at base-1 of Q3. Set the speed switch to 67 wpm and adjust the 67-wpm pot for 100.0 Hz. Capacitor C1 may require some padding to get the 60- and 67-wpm clock frequencies to fall within the range of their respective pots. After this, the pots for 75 and 100 wpm may be adjusted for their respective clock frequencies.

If an oscilloscope or counter is not available, the receiving clock can be adjusted by printing a teleprinter test signal of the desired speed. The test signal should be at machine speed. At first the printer may print garbage because the

clock is too far out. Carefully adjust the appropriate pot until the printer prints correctly. Find the minimum and maximum points of the pot that give good copy, then carefully center the pot between these two points. (Note: Neither control transistor's base should be grounded with this method.)

If a frequency counter isn't used to adjust the output clock frequency, the adjustment should be refined by making range measurements on the printer. Readjust the output clock frequency for maximum printer range while receiving a test signal.

troubleshooting

Be sure the wiring is correct. It's sometimes helpful to have a friend double-check the wiring — you may consistently overlook a mistake.

The unijunction oscillators should free-run whenever the base of the associated control transistor is grounded. If not, check the base-2 supply voltage. Check the voltage across the timing capacitor; it will resemble a sawtooth waveform when the oscillator is running. The oscillator may not be working due to an open charge path to the capacitor, or the capacitor may be leaky. Check the control transistor by lifting its collector lead from the unijunction emitter.

If the speed converter works only partially chances are one or both clock oscillator frequencies are not correct. Recheck them. Also check the input waveforms — the output of Q1 should be no more than 0.3 volt for a space and greater than 2.5 volts for a mark.

Is the regen output working correctly? If so, the receiving portion of the speed converter is working correctly. If not, check the various logic signals with an oscilloscope.

If the trouble appears to be in the output section, check to see if the clock comes on at all right after a character has been received. If the input of the speed converter is spacing, the speed-converted output should also be spacing (running open). If not, the last stage of the shift register is not being preset or its ($\bar{Q}$) output is not connected to U8. The clock should definitely run when the last stage is preset.

Check the logic levels to the output (hub driver) transistor.

If certain bits are consistently garbled there may be trouble in loading the output register.

conclusion

Operation of the speed converter has been quite reliable. At the beginning of the project it was feared that the unijunction clocks wouldn't exhibit sufficient long-term stability and that crystal clocks with dividers would be in order. Such is not the case — the unijunction clocks have been quite stable with precision resistors and mylar capacitors in the timing networks.

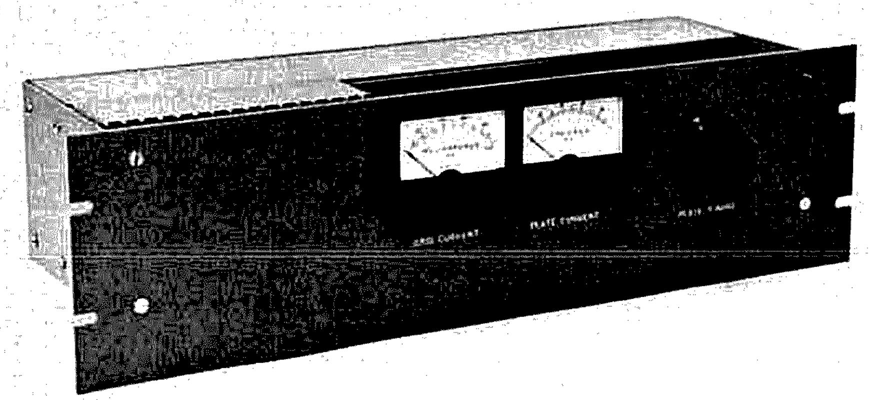
Printing accuracy of the speed-converted machine is not much different than that of a machine geared to the desired speed under adverse receiving conditions. Oftentimes copy is better on the speed-converted machine than on the directly geared machine, probably due to the improved synchronization of the converter's receiving section to the incoming teleprinter signal.

With the possibility of multiple-speed authorization by the FCC for the amateur bands, the speed converter should be a valuable asset to any RTTY station. By suitable transistor switching, the receiving speed of the converter may be changed by control signals from an electronic character recognizer or a stuntbox to allow any of the standard speeds to be programmed into the converter by the received signal. Alternatively, suitable logic could be added to recognize what speed is being received and then change the receiving speed of the converter to that of the incoming signal.

references

1. Petersen, *RTTY Journal*, March, 1969.
2. Haynes, "Standard System for Simple Station Control," *RTTY Journal*, June, 1971.

ham radio



high-power linear amplifier for 220 MHz

This high-performance kilowatt amplifier for the 1¼-meter band uses a new ceramic/metal triode, the Eimac 8874

Robert I. Sutherland, W6UOV, Eimac Division of Varian, San Carlos, California

The Eimac 8874 is a new ceramic/metal, high-mu triode rated for use up to 500 MHz. Operation of this tube in separate 144, 220 and 432 MHz amplifiers has proved that it is an excellent performer up to 450 MHz. This article describes a 220-MHz linear amplifier that uses the 8874.

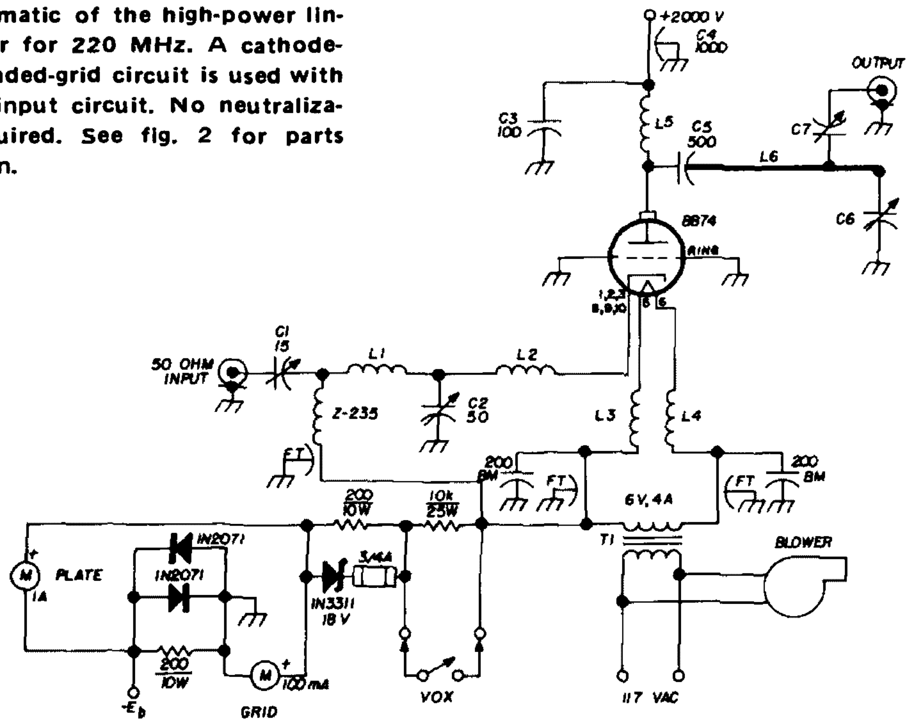
The 8874 has good division between plate and grid current and low intermodulation distortion. The tube has a plate-dissipation rating of 400 watts and mu of approximately 160. The cathode is indirectly heated, and filament requirements are 6.3 volts at 3.2 amperes.

The 220-MHz amplifier described in this article is designed for the serious DXer who works tropo-scatter as well as for the very serious experimenter who is trying to work moonbounce or meteor-scatter. The compact amplifier uses a grounded-grid circuit with a half-wave open plate line and lumped T-network cathode input circuit. The amplifier is not touchy in operation, requires no neutralization and is completely stable and free of parasitics.

The amplifier is intended for 50 per cent duty operation at the 1000-watt dc input level and can develop 1000 watts

PEP input for ssb operation.* For 1000 watts PEP input the plate voltage should be 2000 volts. The plate current under key-down conditions will be 500 milliamperes. Without speech processing the plate current for one kilowatt PEP input 27 pF). The network consists of one series capacitor and two series inductors with one shunt capacitance. Electrically, the network has two inductive reactances in series with a shunt capacitive reactance. The variable capacitor, C1, is in

fig. 1. Schematic of the high-power linear amplifier for 220 MHz. A cathode-driven grounded-grid circuit is used with a T-match input circuit. No neutralization is required. See fig. 2 for parts identification.



under voice conditions will be about 200 milliamperes. Do not *talk* the amplifier up to 500 milliamperes on ssb! If you do, the amplifier will be overdriven and will cause splatter and distortion. The amplifier will deliver 580 watts output; drive power will be 29 watts. Stage gain is 13 decibels at an amplifier efficiency of 58 per cent.

input circuit

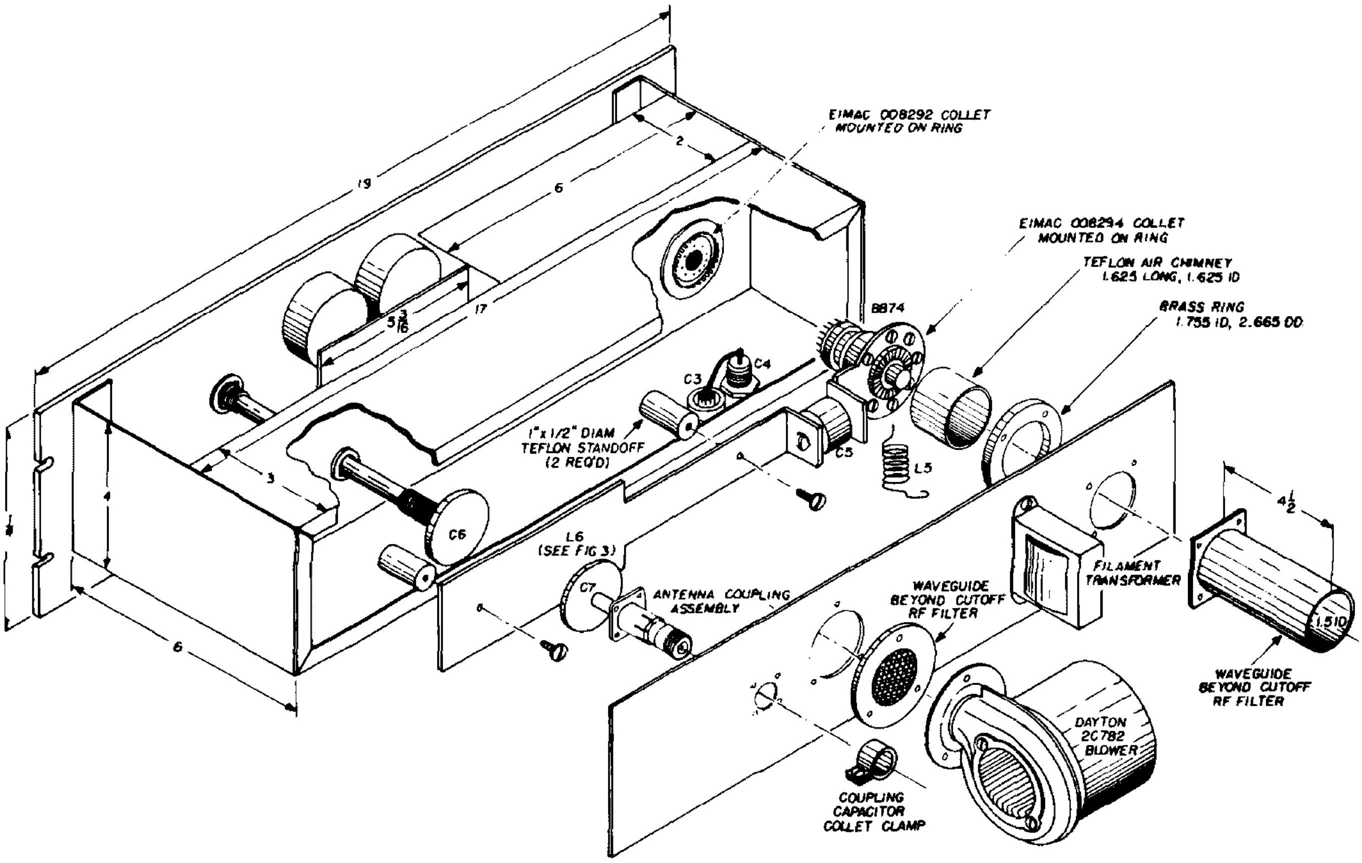
The cathode input matching circuit is a T-network which matches a 50-ohm termination to the input impedance of the tube (about 94 ohms in *parallel* with

*For a discussion of the ratings for this type of operation read the article, "Intermittent Voice Operation of Power Tubes," in *ham radio*, January, 1971, page 24.

series with L1 (see fig. 1) to allow the input inductive reactance to be varied. L1 is larger than the network calculations call for; by placing a variable capacitor in series with L1 the correct value of inductive reactance can be obtained by varying

table 1. Performance data. The conditions most suitable for amateur ssb at 1000 watts PEP and 100 watts cw are as follows:

Plate voltage	2000	V
Plate current (idling)	20	mA
Plate current (single tone)	500	mA
Grid voltage	-12	V
Grid current (single tone)	73	mA
Power input	1000	W
Power output	580	W
Efficiency	58	%
Drive power	29	W
Power gain	13	dB



C1	15-pF variable (Hammarlund MAPC-15)
C2	50-pF variable (Hammarlund MAPC 50)
C3	1000 pF, 2500 volts (Erie 1270-010)
C4	100 pF, 5000 volts (Centralab 850S-100)
C5	500 pF (Centralab 858S-500)
C6	1.75" diameter disc approximately 0.4" from plate line.
C7	1.50" diameter disc approximately 0.375" from plate line
L1	0.8 μH, 4 turns no. 14, 5/16" diameter, ½" long
L2	0.03 μH, 1 turn no. 14, ½" diameter, ½" long
L3,L4,	10 turns no. 16 Formvar, 5/16" diameter
L5	10 turns no. 14, 3/8" diameter, 7/8" long
L6	see fig. 3
T1	6 volt, 4 amp filament transformer (Stancor P6376)

Blower is Dayton 2C782

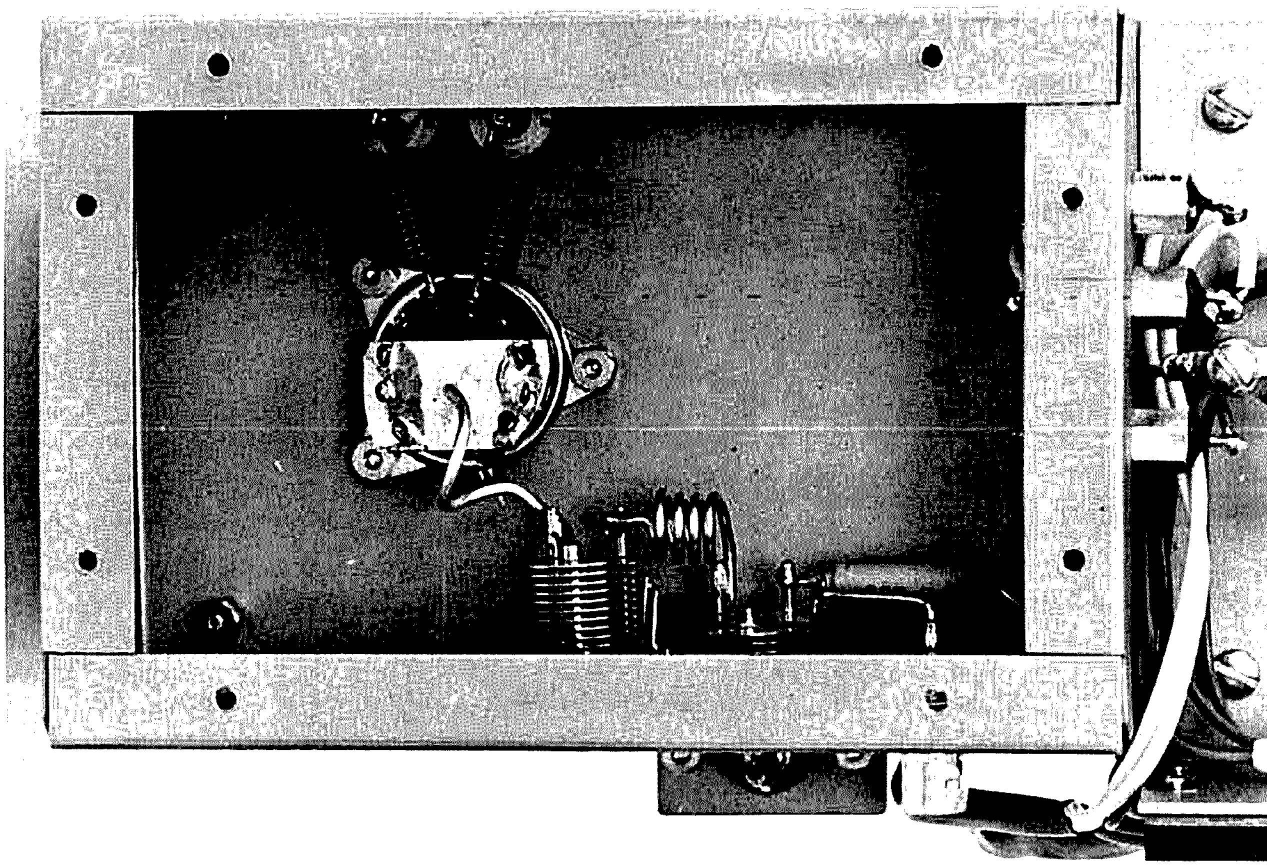
FT are 500-pF feedthrough capacitors (Erie type 327)

fig. 2. Exploded view of the 220-MHz power amplifier.

the capacitor. The two variable capacitors allow the network to cover a wide range of impedance transformations.

The variable capacitor C1 is mounted on the bottom side of the input circuit chassis. The shunt capacitor, C2, is also mounted on this side of the chassis to allow easy adjustment from the bottom of the amplifier. Front panel access to these two controls did not seem desirable since these adjustments don't require attention very often.

The layout of input circuit components is seen in the photographs. The



Cathode input circuit. Cathode terminals are interconnected with small copper plate. Filament chokes L3 and L4 are at top. Cathode input matching circuit is at the lower right inside the box.

input network is in the chassis box at the left. Capacitor C1 is the smaller capacitor adjacent to the rf input coax fitting. L1 is the coil running between the adjacent variable capacitors, C1 and C2. The one-turn inductance between the center of the tube socket and the variable capacitor C2 is L2. This coil is terminated in the center of the copper plate connecting all cathode leads together to keep the path length to all cathode leads the same.

The photograph of the socket assembly shows the control-grid collet assembly which is used to ground the control grid

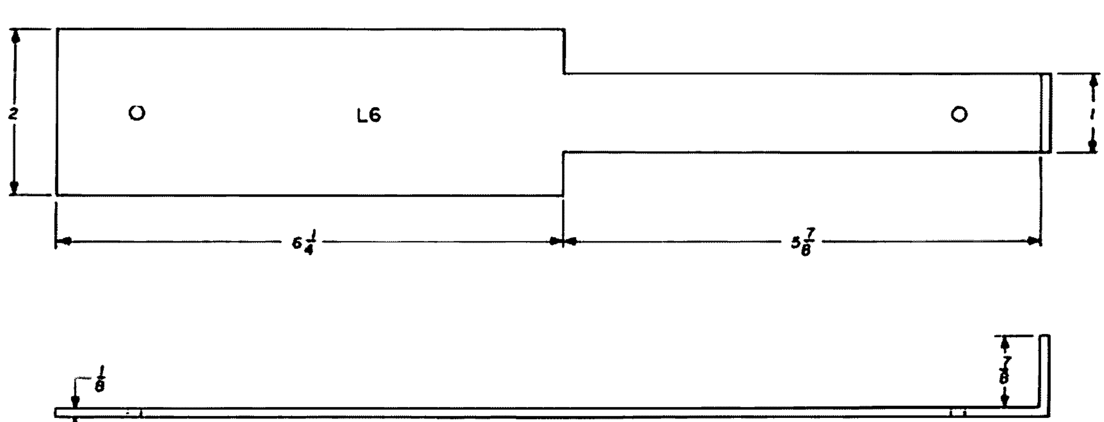
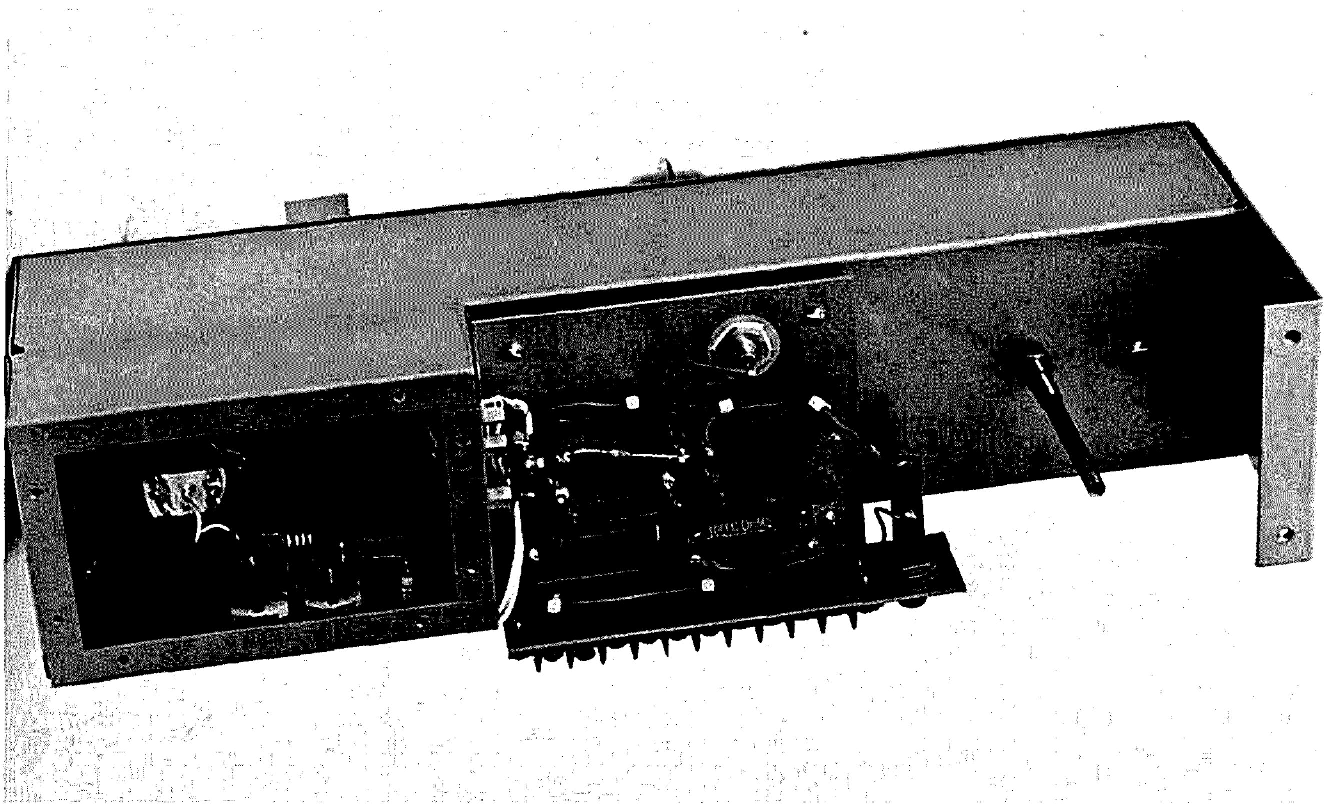


fig. 3. Plate line for the 220 MHz amplifier.

and to mount the socket. The control-grid contact is an Eimac 008292 collet soldered to a 1/16-inch thick copper ring, two inches in diameter with a hole 1-7/16 inches in diameter in the center. The three mounting screws are 4/40 captive stainless studs, 1/4-inch long, and are positioned to line up with the holes in the Johnson socket (Johnson 124-311-100).

Standby plate current of the 8874 is reduced to a very low value by use of a 10k, 25-watt cathode resistor which is shorted out when the amplifier vox circuit is energized, permitting the tube to operate in normal fashion. A 200-ohm safety resistor insures that the negative power circuit of the amplifier does not rise above ground potential if the positive terminal of the plate supply is accidentally grounded. A second safety resistor is placed across the zener diode to provide a load for the zener; it also prevents the cathode potential from rising if the zener should accidentally burn open.

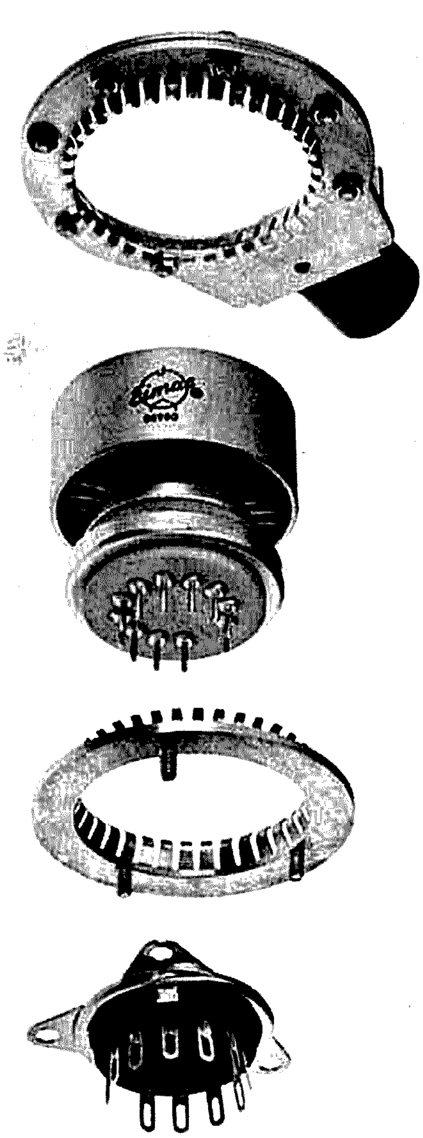


Amplifier with front panel removed. Cathode input circuit is in box at left with tuning capacitors C1 and C2 mounted on bottom. The 8874 tube socket, with cathode terminals interconnected by a small copper plate, is mounted to the rear of the box. At center are the resistors, zener diode, 1N2071 diodes and the zener protection fuse. The shaft at the right is the plate circuit tuning control.

The amplifier grid (fig. 1) is operated at dc ground; the grid ring at the base of the 8874 provides a low-inductance path between the grid element and the chassis. Plate and grid currents are measured in the cathode return lead with a 12-volt, 50-watt zener diode in series with the negative return to set the desired value of idling current. Two additional diodes are shunted across the meter circuit to protect the movements against destructive overload.

plate circuit

The plate circuit of the amplifier is a half-wave open transmission line. A quarter-wave line could have been used with a small improvement in bandwidth



Exploded view of the tube socket, grid collet, tube and anode connector. The anode assembly is made of two copper rings which encircle the tube; the rings are clamped together with the collet in between. Bottom ring has a flange which is attached to plate blocking capacitor C5.

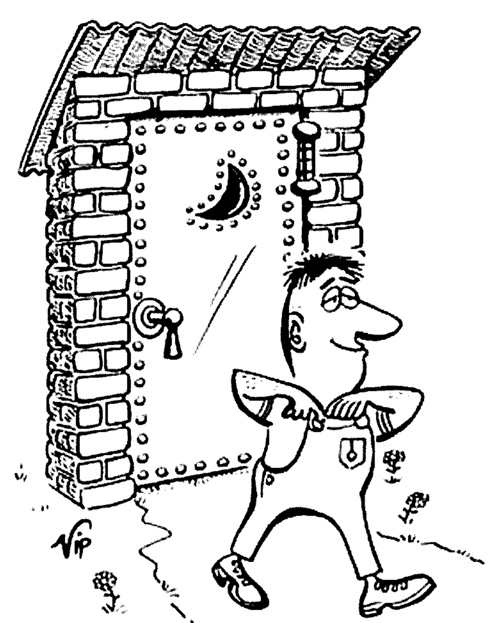
and efficiency. However, the size of the plate circuit would be very small and more difficult to work with. A quarter-wave line would terminate approximately at the first mounting screw for the plate-line mount.

The half-wave line was made wide at the open end to allow sufficient area so that enough capacitance could be obtained in the tuning and loading capacitors without reducing the plate-to-plate spacing which would degrade the voltage hold-off capability. A low-impedance half-wave line was too long to fit into the 17 inch chassis. Therefore, a shorter line was narrowed at the high current point to increase the inductance and make the line electrically longer.

The plate line is resonated by C6 while the loading is adjusted by C7. These two adjustments will interact to a certain extent and therefore the proper operating point must be determined by adjusting both controls several times.

A type-N coaxial fitting is connected to the moveable disc of the coupling capacitor. The fitting is centered in a special tubular assembly which permits the whole connector to slide in and out of the chassis mounting fixture; this allows the variable disc of the coupling capacitor to move with respect to the fixed plate mounted on the tube anode clamp. When the final loading adjustment has been set the sliding fitting is clamped by a fixture similar to the slider on a variable wire-wound resistor.

The disc is mounted on a threaded shaft which moves in and out through the threaded bushing on the front panel. To avoid jumpy tuning a fine thread was used. An alternative would be to give added support to the shaft by mounting a threaded nut on a strap placed ½- to ¾-inch from the sub-panel. This would



Metal/ceramic construction now in Eimac zero-bias triodes

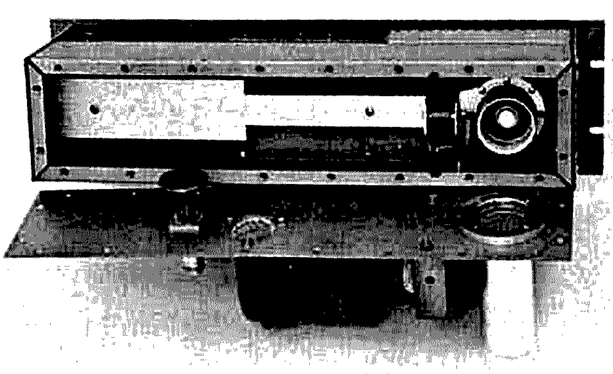
provide two supports for the threaded shaft.

The plate contact assembly is made from two copper rings and an Eimac 008294 collet clamped together with 4-40 brass machine screws. One of the rings in the clamp has a flange to provide a mounting bracket for the plate blocking capacitor, C5.

The Centralab capacitor (C3) mounted in parallel with the feedthrough, C4, was necessary to remove rf from the plate voltage lead outside the enclosure. The feedthrough capacitor did not do the job by itself.

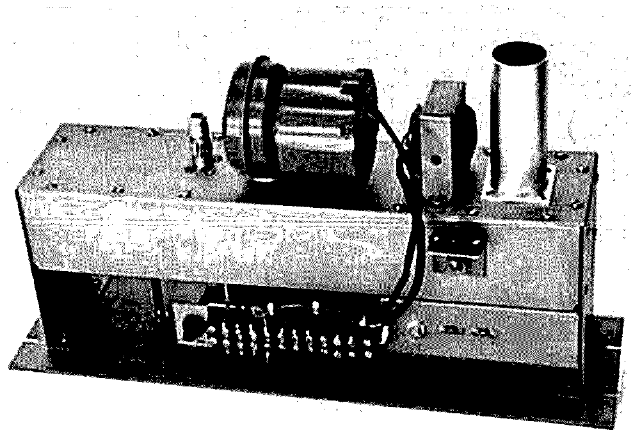
layout and design

The amplifier is built with two standard aluminum chassis. The plate compartment is made from a 3 x 4 x 17-inch chassis; the grid enclosure is made from a 2 x 4 x 6-inch chassis assembly that is attached to a standard 5¼ x 19-inch relay-rack panel. The front panel also supports the grid and plate current meters. Just behind the panel and next to the grid compartment (which is mounted at one end of the longer chassis) is a flat aluminum plate that supports the three large resistors, the zener diode, terminal strip and the fuse in series with the zener.



Anode circuit. Tube is at right, with plate line extending to the left. The disc at the left end of the cover plate is the load coupling capacitor C7. A Teflon chimney for the 8874 sits on top the tube anode and protrudes out of the waveguide-beyond-cutoff air pipe in the cover plate. Note the rf filter in the exhaust port of the blower.

The pictorial drawing in fig. 2 shows how the whole chassis assembly is put together. The back panel of the plate compartment is used to make an rf shield, an air-tight enclosure, and to serve as the mounting deck for the blower, the filament transformer, the rf output connector and the exhaust port for the cooling system. The centrifugal blower forces cooling air into the plate compartment and the air escapes through the anode cooling fins of the 8874, the Teflon air chimney and finally, out through a waveguide-beyond-cutoff air pipe.

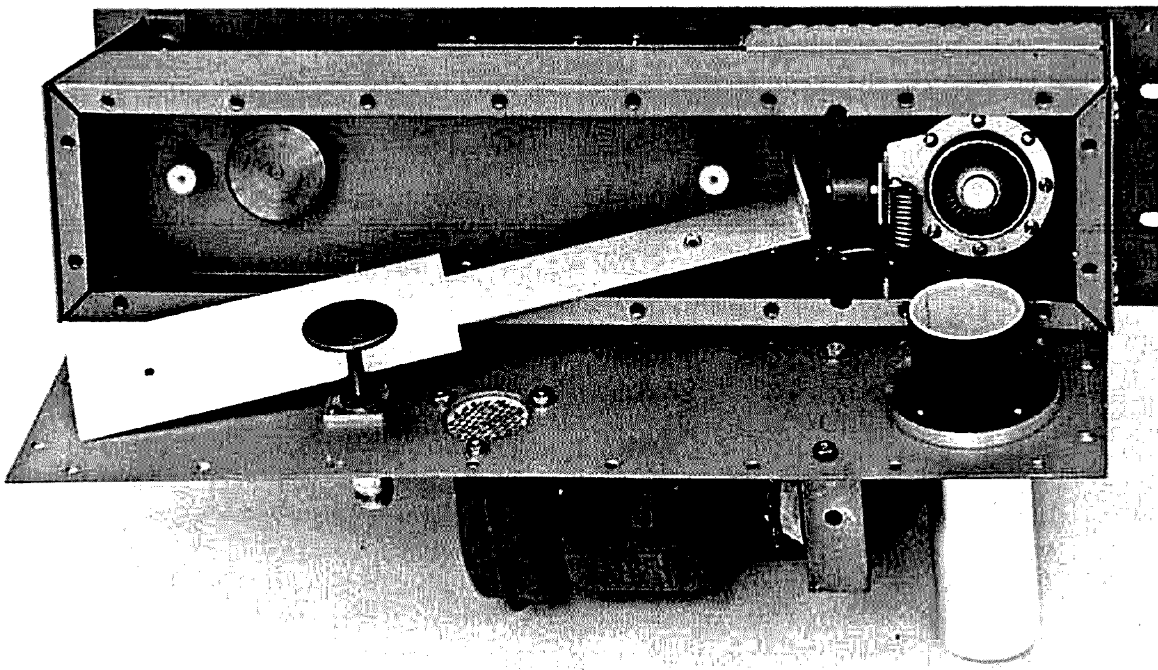


Rear view of 220-MHz amplifier. The two adjustments for the cathode circuit and input coaxial receptacle are at lower right. Type-N output connector is to left of the blower. Filament transformer and air duct are at the right-hand end.

The cooling-air inlet hole is also shielded from rf using the waveguide-beyond-cutoff technique. In this case a piece of copper honeycomb similar to a radiator core was soldered into the center of a ring and the assembly mounted between the blower outlet and the back plate of the chassis.

operation

Amplifier operation is completely stable with no parasitics. The unit tunes smoothly and the plate current dip occurs at the same time as the power output peaks. As with all grounded-grid amplifiers, excitation should never be applied



Anode compartment with plate line removed. Line has been displaced from its two Teflon support pillars to show the moveable disc of the plate tuning capacitor C6.

when plate voltage is removed from the amplifier.

The first step is to grid-dip the input and output circuits to near resonance with the 8874 in the socket. An swr meter should be placed in series with the input line and the input network adjusted for lowest vswr on the line from the exciter.

Tuning and loading follows the same sequence as any grounded-grid amplifier. Connect a swr indicator and load to the output of the amplifier and apply a small amount of rf drive. Quickly tune the plate circuit to resonance.

The cathode circuit should now be resonated. The vswr measured between the exciter and the amplifier will not necessarily be optimum. Final adjustment on the cathode circuit for minimum vswr should be done at full power level because the input impedance of a cathode-driven amplifier is a function of the plate current of the tube.

Increase the rf drive in small increments along with the output coupling until the desired power level is reached. By simultaneously adjusting the drive and loading it will be possible to attain the operating conditions given in the performance chart in table 1. Always tune for maximum plate efficiency, that is, maximum output power for minimum input power. It is quite easy to load heavily and underdrive to get the desired power input, but power output will be down if this is done.

My thanks to Dick Rasor, WA6NXB, for his help in adjusting and determining the operating conditions for this amplifier.

ham radio

evaluating semiconductor diodes

Here are three simple tests for determining important diode characteristics: germanium or silicon, peak-inverse voltage and maximum current

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Semiconductor diodes are widely used by radio amateurs. Fortunately, surplus diodes are both plentiful and inexpensive. However, not all are marked, and even some that are marked carry *house* identifications that are difficult if not impossible to check for characteristics. This article describes some checks that are easily made with equipment available in the average experimenter's workshop to determine many of the characteristics you need to know to put the diodes into service.

silicon or germanium?

First you'd like to know whether the unidentified diode is made of germanium or silicon. Since selenium and copper-oxide diodes are not often found they will not be discussed here. Here's a quick and easy test. Note the circuit in fig. 1; it calls for a source of variable direct-current voltage, a couple of meters and a resistor. You'll need only a few volts from the supply, the meters can be multimeters of an inexpensive variety, and the resistor is quite noncritical, as it serves as a current limiter and must handle only a few milliamperes. The voltage meter should have a scale that lets you easily tell the difference between 0.4 and 0.7 volt. The other meter should be a low-range milliammeter, preferably one with rather high internal resistance. Note that the diode is connected for forward conductance.

To determine the nature of the diode, slowly advance the voltage from the variable source until the milliammeter starts an abrupt spurt of current. Stop right there and read the voltage. If it's in the general range of 0.4 volt, you have a germanium diode. If it took around 0.7 volt, it's a silicon diode.

peak-inverse voltage

Now that you know what type diode you have, your next thought may relate to just what peak-inverse voltage you can impose on the diode and still keep it viable. For this check, you'll need

another variable-voltage power source with a much higher upper potential, one that exceeds any expected PIV. The test setup is shown in **fig. 2.** Note that this time you put the diode in backwards; that is, you connect it in the normally nonconducting direction.

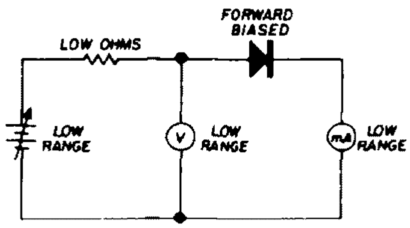


fig. 1. To determine whether the diode is silicon or germanium, put the unknown diode in this circuit, adjust the power supply until the milliameter starts to show current, and read the voltmeter. Voltage in the vicinity of 0.4 volts indicates a germanium device; 0.7 volts, a silicon diode.

The resistor needs to be large enough to limit the current to a few milliamperes; if you're expecting to use over one-thousand volts it's best to use several resistors in series to discourage voltage breakdown. The test procedure is the same as before: increase the voltage slowly until the milliammeter starts showing current. Stop there, note the voltage, and back off before you destroy the diode.

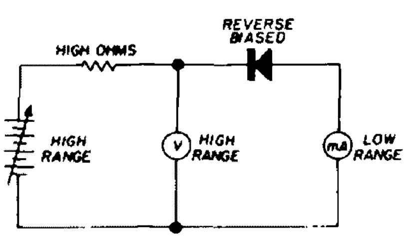


fig. 2. Diode PIV test circuit. Adjust the power supply until the milliameter shows current and read the PIV on the highrange voltmeter.

maximum current

Maximum current-carrying capability is not so easily determined. You can get an approximation, however, with the circuit shown in **fig. 3.** Here again the resistor is for current limiting and must carry the maximum expected current capability of the diode. The variable-voltage source will not need to provide much voltage, but it must be capable of supplying the maximum current you suspect the diode may sustain. The ammeter, of course, also must have a range encompassing the highest current expected.

The check will not give you precise results, as you must make one qualitative evaluation: the heat of the diode. That's how you make the test; you increase the current flow until the diode gets hot. Just how hot is a matter you must decide. If it's a small diode, cooled only by the leads and convection air flow, you'd better stop when it starts to get warm. If, though, it's a husky brute, meant to be

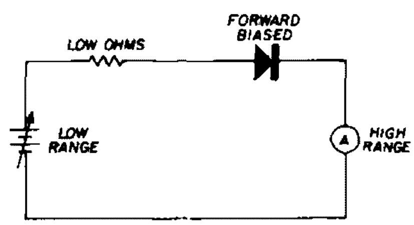


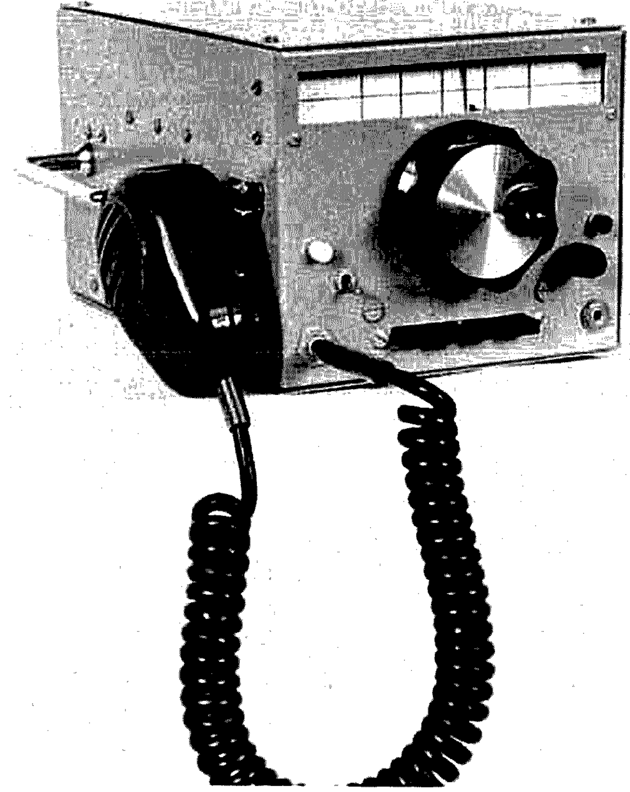
fig. 3. Maximum current-carrying capability of the diode is roughly determined by the amount of heat dissipated by the device in this circuit (see text).

mounted on a big heat sink, you can run it to the finger-burning stage! Just don't get carried away with trying to push the current to the utmost limit, or you may end up with a cooked diode.

Also, on PIV or maximum-current tests, don't check just one diode and assume that all of that type or number will show the same characteristics; the probabilities are that they will not. In PIV, especially, they'll vary over a wide range. And not all germanium or silicon diodes will exhibit a uniform forward-conduction threshold voltage.

These three simple tests, easily made with commonly-available equipment, will give you most of the information you need to put those stray diodes into use in your ham shack.

ham radio



amplitude-modulated two-meter transmitter-receiver

This compact transceiver for two meters features a high-performance mosfet receiving converter, IC i-f and audio and 10-watt rf output

Robert A. Thompson, K1AOB, 60 Buck Thorne Avenue, Riverside, Rhode Island

Riding on the coat tails of the current wave of two-meter fm activity is an increase in two-meter operation in general. SCR-522s and ARC-4s are being dusted off and fired up again; mountain-topping and mobiling have attracted renewed enthusiasm.

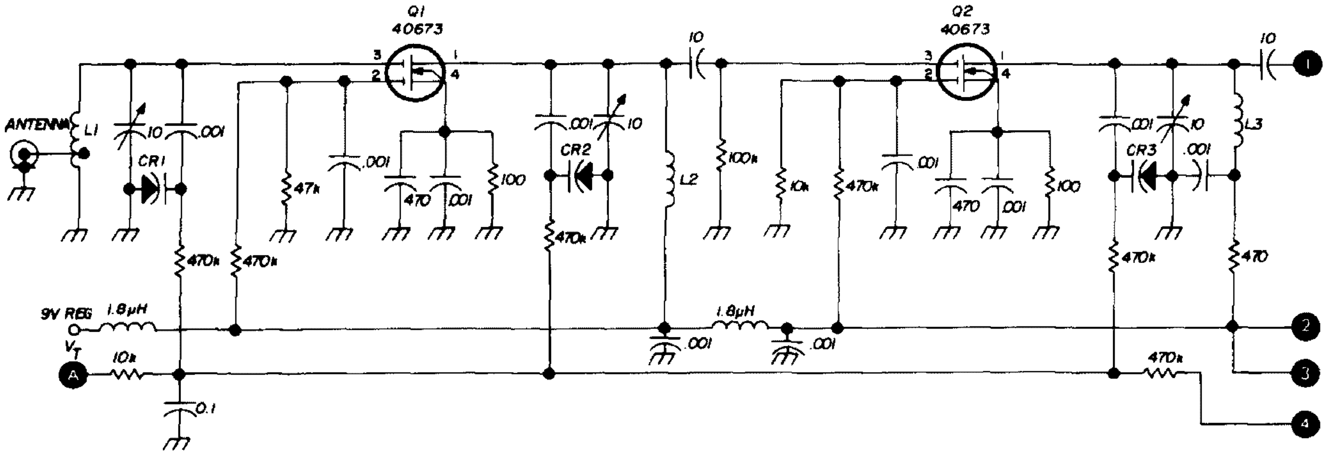
The two-meter a-m transmitter-receiver described here is an attempt to provide a unit suitable for mountain topping and mobiling. The rig serves its purpose admirably and has provided much enjoyment. However, you must remember that this is a minimum-specification rig; it does not contain any luxury features which might be desired by others.

Since advanced vhf constructors seldom duplicate equipment completely this article is presented more as a source of ideas than as a one-for-one construction article. For this reason the unit schematics are shown individually.

performance

With an input of 12.6 volts dc the transmitter will provide 10 watts into a 50-ohm load with average modulation. Slightly greater power output is available with the engine running as the alternator usually increases the voltage to about 13.5 volts. The modulator is capable of producing more than ample audio and must be initially adjusted with the internal input potentiometer to prevent splatter.

The receiver has some unique aspects. For example, the dial has no frequency markings but is divided into four blocks. There are four push-buttons under the



L1	3 turns no. 22, airwound, 1/4" diameter, 3/4" long, tapped 1 turn from ground end	L4	3 turns no 16, airwound, 1/4" diameter, 3/4" long
L2,L3	3 turns no. 22, airwound, 1/4" diameter, 3/4" long	T1,T2,T3	see text

tuning knob. When then the first button is depressed, the receiver tunes from 144 to 145 MHz. Therefore, each block on the dial represents 250 kHz.

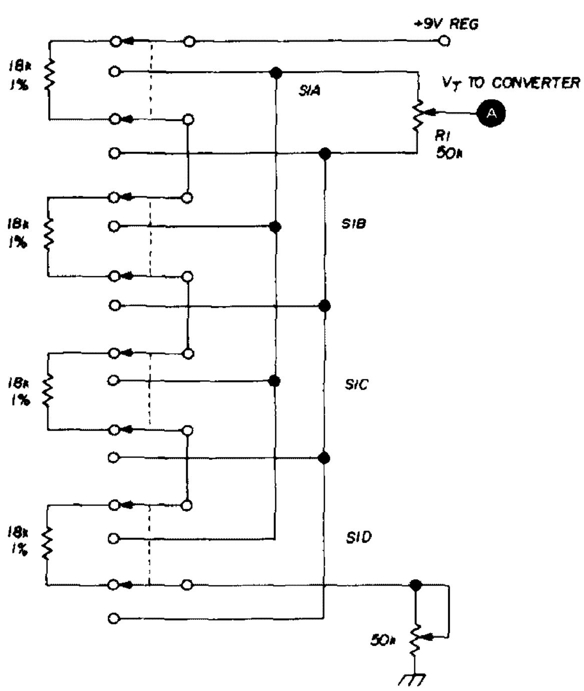


fig. 2. The MHz selector switch; S1 is a 4-station push-button switch with interlock.

These dial segments can be further subdivided for additional accuracy if desired. The push-buttons interlock so that when the second button is depressed the first disengages and the dial now represents frequencies from 145 to 146 MHz.

Since the dial is four inches long this arrangement is equivalent to a dial sixteen inches long! It takes about three and one half turns of the tuning knob to traverse the dial; this is 14 turns for the entire two-meter band.

converter

To tune the converter a variable voltage is used to reverse bias diodes CR1, CR2, CR3 and CR4. (See **fig. 1.**) When any solid state diode is reverse biased its junction exhibits a capacitance effect; the BA-110 diodes are specifically designed for use in tuned circuits.

They are effectively across the tuned circuits of the rf amplifier and local oscillator. This method of tuning frees the builder from the bulky tuning capacitor and the necessity of mechanically coupling it to the tuning knob and dial. With this method you could even mount the converter on the antenna and tune it remotely!

Two stages of rf amplification provide input sensitivity on the order of 0.1 microvolt. RCA 40673 mosfets are used in the rf amplifiers and mixer. These transistors include internal protection against breakdown from static electricity.

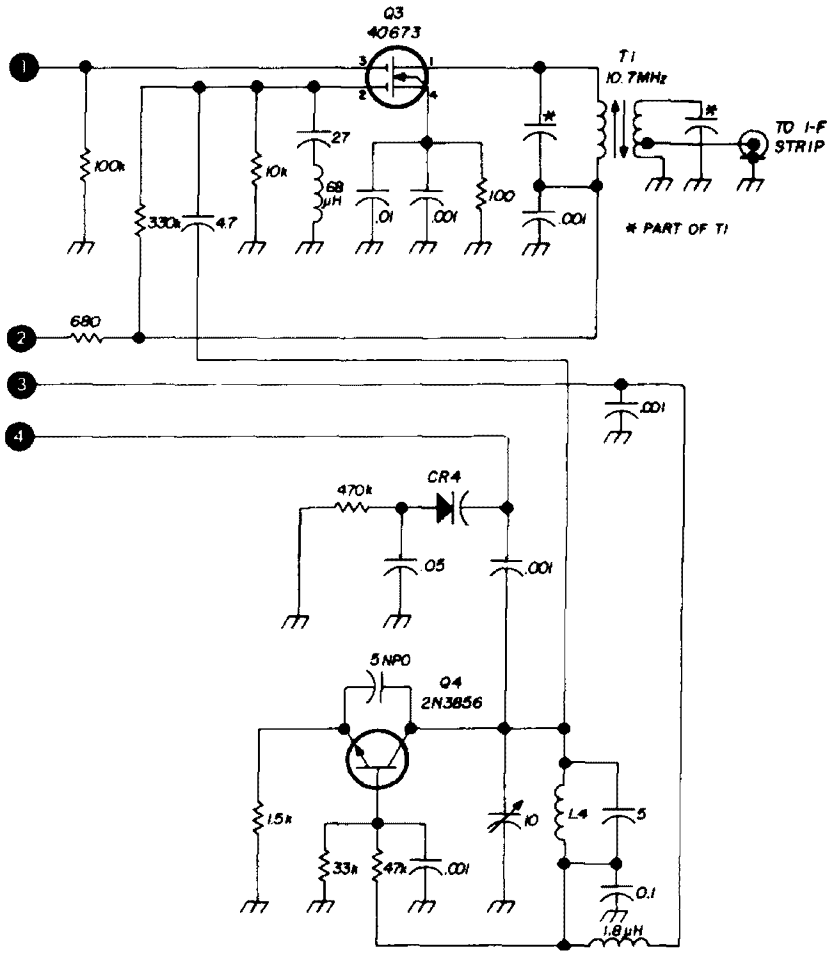


fig. 1. Two-meter receiving converter. All variable capacitors marked 10 pF are 1.5 to 10 pF compression trimmers. Diodes CR1 through CR4 are ITT type BA110 tuning diodes.

Later addition of an fm detector was considered so provision was made to use its output voltage to feed an afc correction circuit. If fm will not be used all circuitry to the left of CR4 may be eliminated.

Double-tuned i-f transformers are preferable to single-tuned units at the mixer output (T1 in fig. 1). J. W. Miller manufactures several types which are satisfactory.

Oscillator coil L4 should be mounted as ruggedly as possible and secured with plastic cement to prevent any vibration. All coils are mounted at right angles to each other so shielding was unnecessary.

i-f strip

Single-tuned transformers are used in the i-f strip (fig. 2) because bandwidth is determined by the crystal filter that preceeds them. The transformers are used merely for interstage coupling. The i-f strip has adequate gain and about 50-dB of agc. Transistor Q5 provides some post-filter amplification while Q6 supplies agc and doubles as a detector.

receiver audio

The detector output is fed to an fet which drives a General Electric PA237 IC power amplifier (fig. 3). A problem with oscillation in the audio circuit was solved by the 8-ohm resistor across the speaker; this is made up of two 16-ohm, ½-watt resistors in parallel.

Maximum available audio is reduced slightly, but oscillations are eliminated, and the IC is protected if the speaker is disconnected with power applied.

The speaker was originally mounted internally. However, speaker vibration produced frequency modulation of the converter local oscillator so the speaker connection was brought out to a jack on the front panel.

voltage regulator

Since a voltage varied by a potentiometer is used to tune the receiver, that

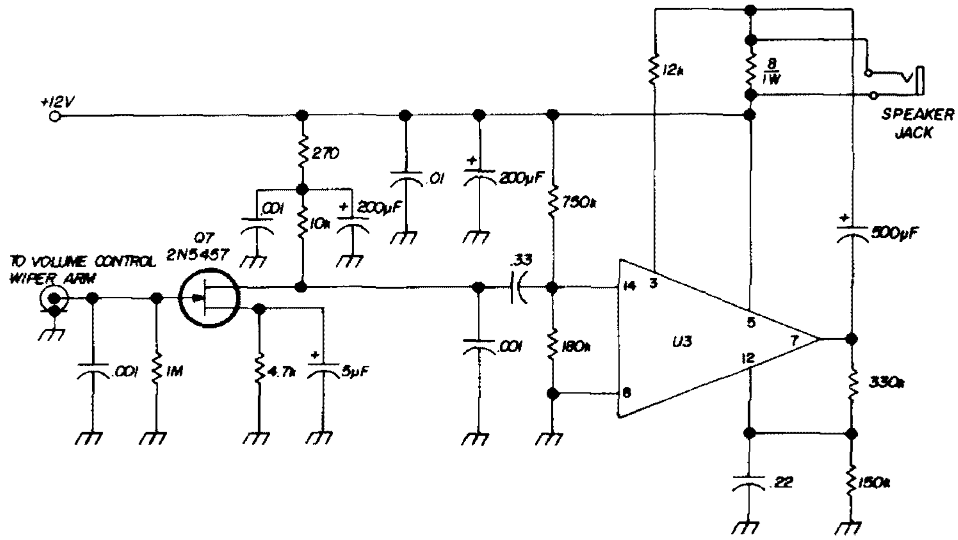


fig. 4. Receiver audio amplifier. Q7 is a 2N5457 or MPF103; U3 is a General-Electric PA237 IC audio power amplifier.

voltage must be very stable. Considering the voltage variation across the average automobile battery under normal driving conditions, this is no easy feat.

The voltage regulator in fig. 5 maintains a constant output plus or minus a few millivolts with an input voltage swing from eleven to twenty volts. The resulting

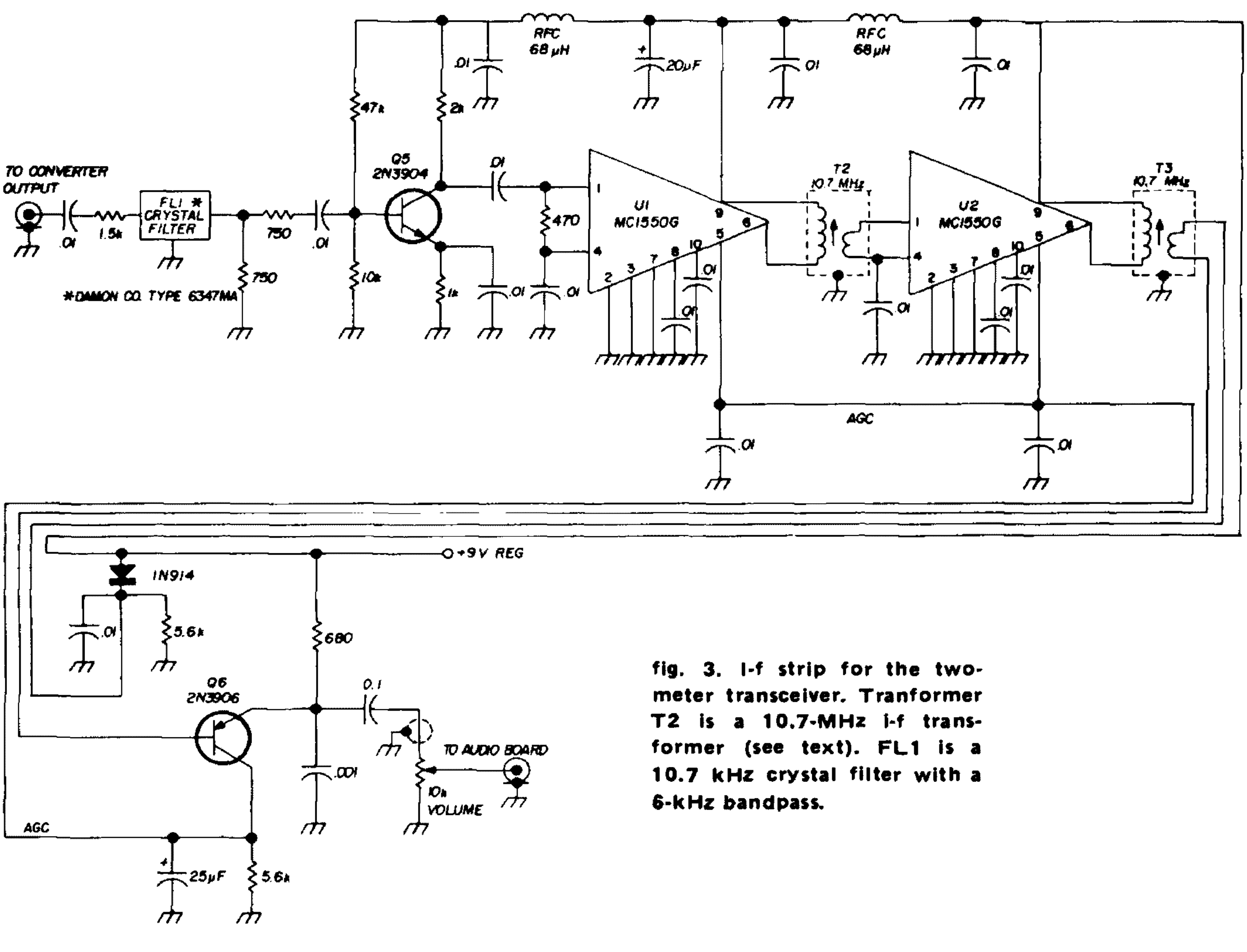


fig. 3. I-f strip for the two-meter transceiver. Tranformer T2 is a 10.7-MHz i-f transformer (see text). FL1 is a 10.7 kHz crystal filter with a 6-kHz bandpass.

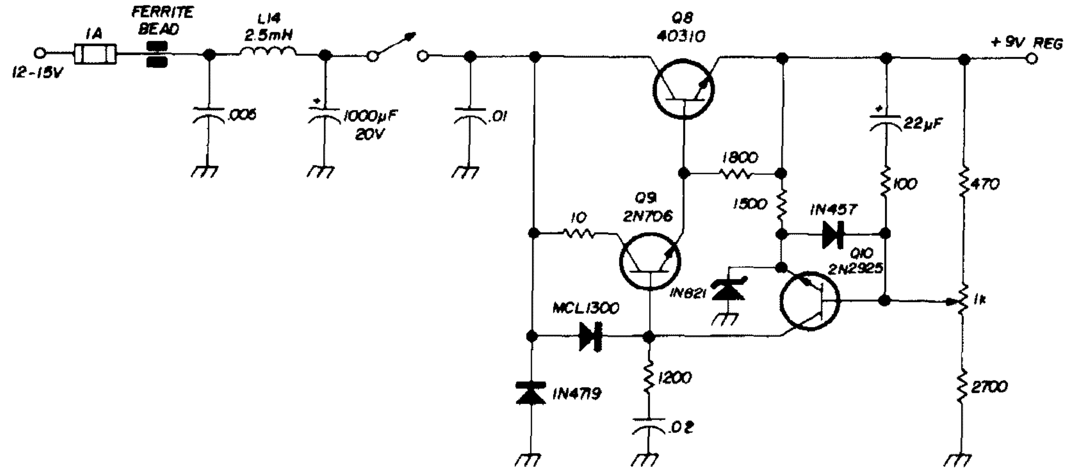


fig. 5. Regulated power supply provides constant 9-volt output with varying 12- to 15-volt input. Choke L14 is a Stancor TC-1.

regulated 9 volts is used only for the i-f strip and the converter. Everything else is fed from the 12-volt input to the unit.

The regulator is capable of much more current than is required in this application, and the circuit might be somewhat tion-level control is mounted on the board; it is adjusted only once during initial tuneup. Heat sinks must be provided for Q12 and Q13.

The oscillator-driver strip is capable of producing three watts into 50 ohms and

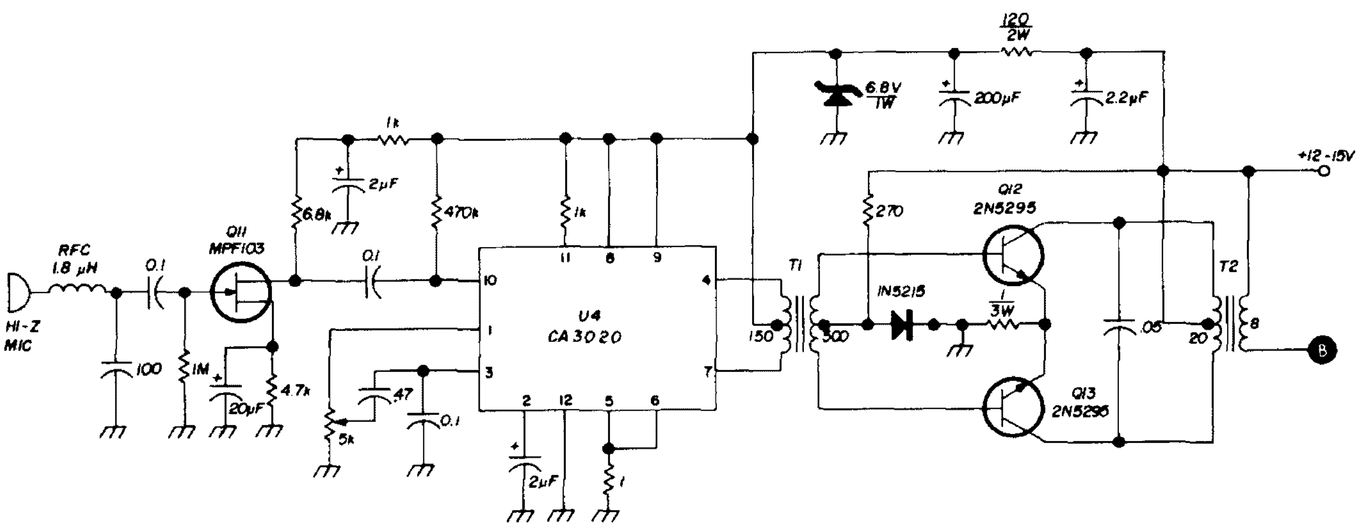


fig. 6. Amplitude modulator for the two-meter transceiver. Transformer T1, 150:500 ohms, is an Argonne AR163 (Lafayette); transformer T2, 20:8 ohms, is a Stancor TA-12.

simplified. Perhaps one of the new IC regulators might suffice.

transmitter

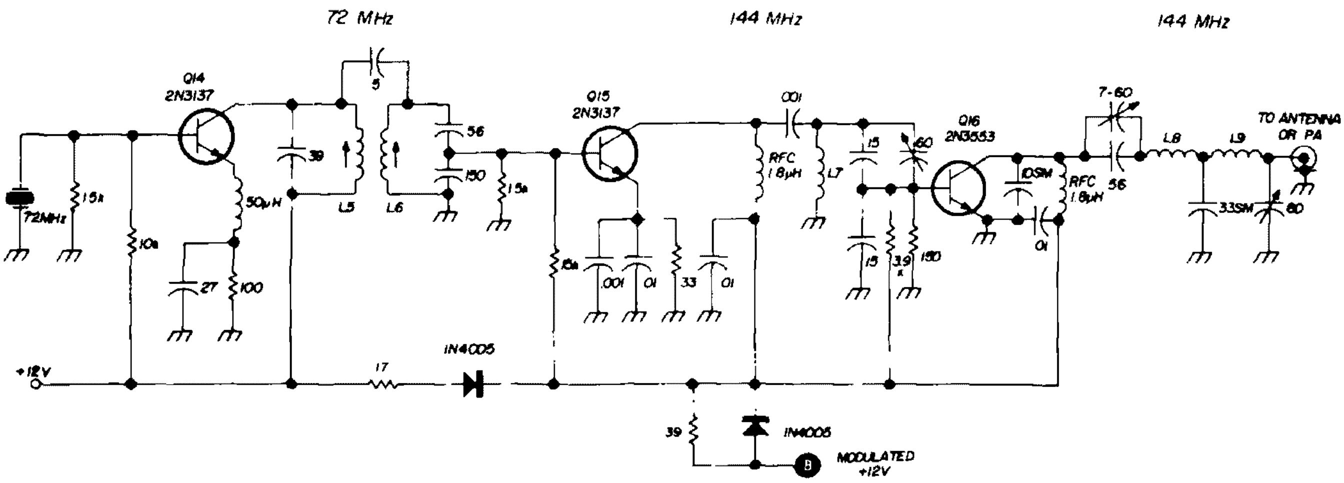
In the modulator a high-impedance microphone feeds an fet which drives a CA3020 integrated circuit, followed by transistors Q12 and Q13. The 5k modula- can be used without the amplifier. An overtone crystal oscillator (Q14) supplies 72-MHz drive to Q15 via bandpass coupling consisting of L5 and L6. Transistor Q15 doubles to 144 MHz and drives Q16 straight through on two meters. The 2N3137 works particularly well in this circuit. Other rf transistors tried for Q14

and Q15 resulted in lower output power.

The final power amplifier, Q17, uses a transistor rated for lower frequencies. Higher frequency types were tried initially, but oscillations were difficult to

ley may be used as these are usually within 2%.

The value of the divider resistors is determined by the number of turns of the 10-turn pot that are required to traverse



L5,L6	3 turns no. 22 on ¼" form, ½" long
L7,L9	4 turns no. 22, airwound, 5/16" ID, ½" long
L8	6 turns no. 22, airwound, 5/16" ID, ½" long

fig. 7. Oscillator-driver for the two-meter transmitter.

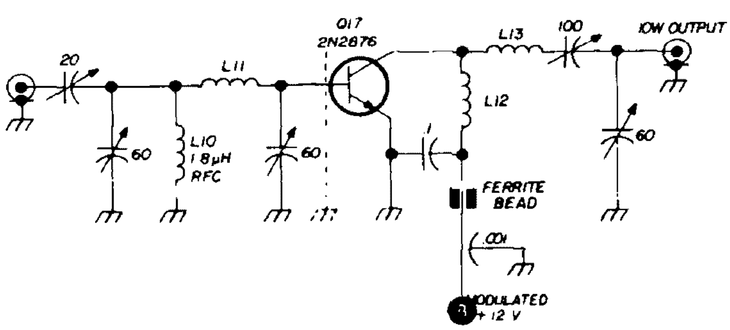
control, and tolerance for load mismatch was nil. The 2N2876 exhibited none of these problems.

mhz selector

Each push-button in the MHz selector switch is a dpdt unit. The divider resistors are all 18,000 ohms, 1%. However, the 5% resistors manufactured by Allen Brad-

the dial once, the voltage-capacitance curves of the tuning diodes and the amount of frequency coverage per switch.

For different dial lengths and/or tuning diodes I would recommend an experimental approach rather than a mathematical one due to the number of variables involved. The 50k pot sets the lower voltage limit of the divider network



L10	1.8 μH RFC
L11	7 turns no. 24, airwound, ¼" ID
L12,L13	4 turns no. 24, airwound, ¼" ID

fig. 8. Ten-watt power amplifier for the two-meter transceiver.

and hence, the low-voltage end of the tuning range.

Depression of a push-button puts the 10-turn pot in series with three of the 18k resistors that make up the voltage divider. If desired the entire band can be tuned in one sweep of the dial by eliminating the push-button switch and divider resistors. This will, however, reduce bandspread by a factor of four.

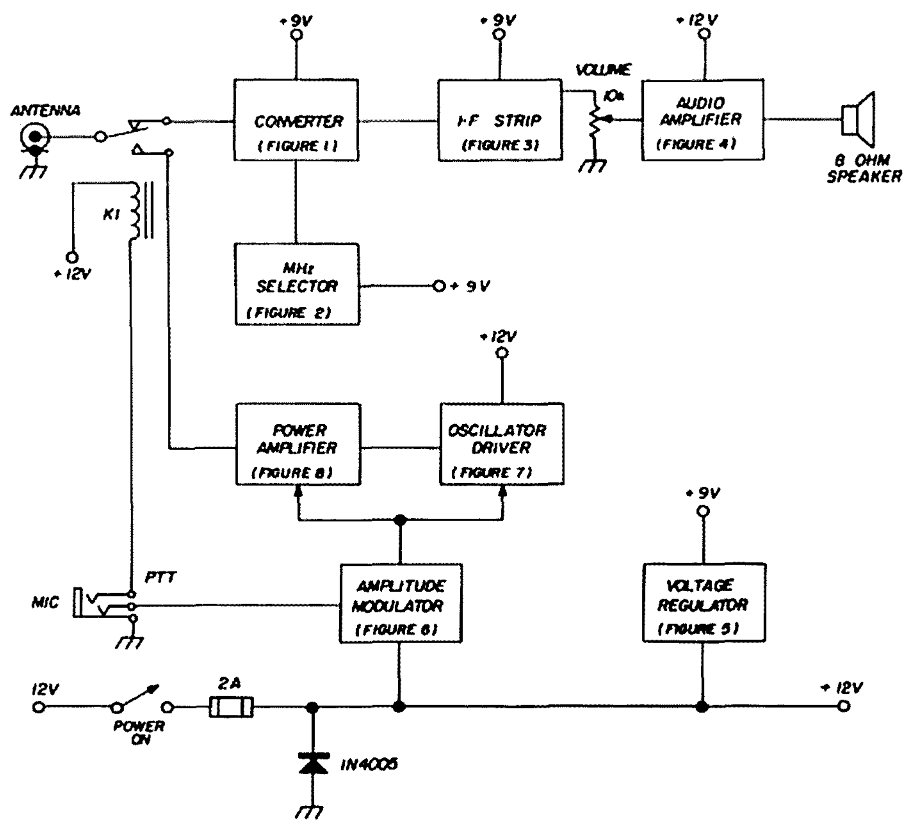
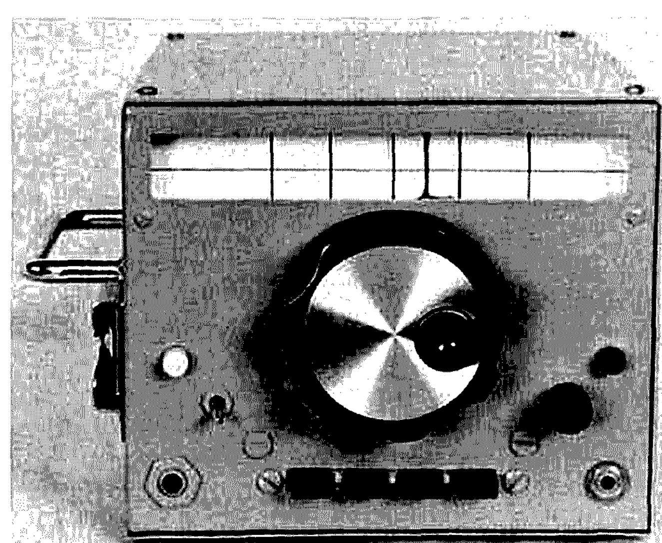


fig. 9. Interconnection of each of the modules used in the two-meter transceiver. K1 is a 12-volt coaxial relay. K2 removes power from the converter and i-f strip during transmit.

construction

The rig is built in a 5 x 6 x 9-inch aluminum utility box. An aluminum panel is placed across the inside of the box about 2 inches down from the top. The transmitter and modulator printed-circuit boards are mounted on this panel with 3/8-inch threaded standoffs.

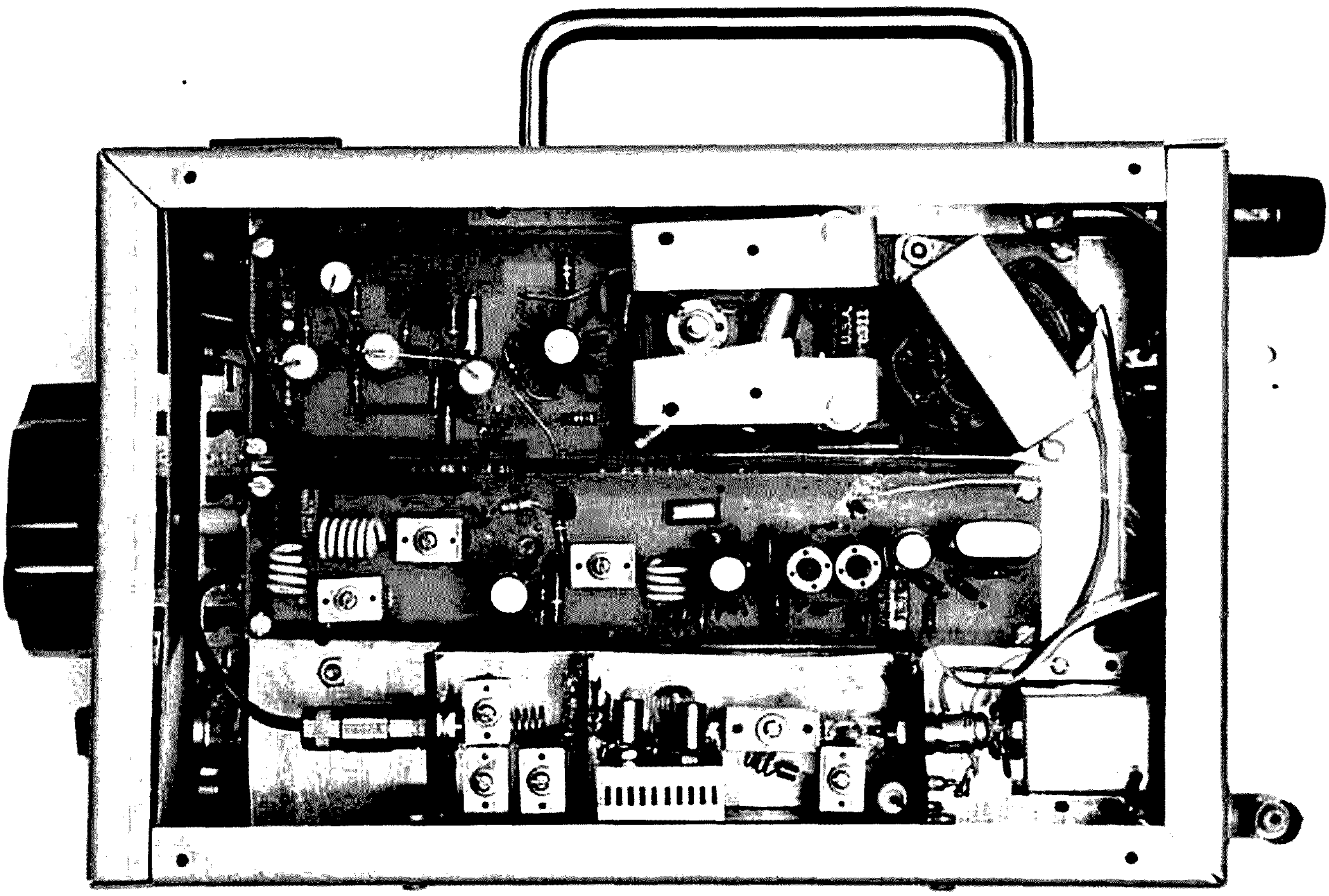
The front end, i-f strip, audio amplifier and voltage regulator are mounted around the sides and the rear wall of the enclosure. This arrangement lends itself to adjustments and ease of construction. A Jones plug for power, a fuse holder and a BNC antenna jack are mounted on the



Front panel of the two-meter transceiver.

rear panel. On the front panel are the tuning knob, a power switch, the volume control, the push-button MHz selector switch, indicator lights and jacks for the speaker and microphone.

rf and mixer transistors, and a plastic socket was used for the oscillator. Nylon or plastic sockets may be acceptable for the rf and mixer stages, but the light brown Bakelite socket should be avoided



The modular construction style is shown in this detail photo of the rig with the bottom cover removed. Note the power amplifier wiring at the bottom.

at these frequencies.

Terminals are silver pins pushed through Teflon inserts. These were used mainly to conserve space; any other type may be used.

The i-f cans were laid on their sides and soldered to the board. IC sockets were also laid on their sides and epoxied to the board. This method of construction was previously described.[5] The audio amplifier is built in the same manner. The voltage regulator is mounted on a conventional printed-circuit board.

The shaft that connects the tuning knob to the 10-turn pot should be non-

Printed-circuit boards were used for the modulator and oscillator-driver. The power amplifier is wired on double clad 1/8-inch thick board but is made with conventional chassis-type construction. There is a shield partition between the input circuitry and the base of the power transistor. The 2N2876 is mounted on a 1 x 1 x ½-inch block of aluminum which serves as a heat sink.

The tunable converter was also built on a piece of double-sided printed-circuit board. Holes were drilled and transistor sockets mounted in the conventional manner; Teflon sockets were used for the

metallic. A microphone clip is provided on one side of the box, and behind it is a handle. The side mounting does not interfere with under-the-dash mounting arrangements and still provides utility.

The slide-rule dial incorporates a length of dial cord looped over the tuning shaft twice and around two small pulleys at the bottom corners of the dial opening. The dial cord is tied to opposite ends of a small spring which provides the necessary tension. A small piece of plastic-covered solid copper wire is looped around the spring and held with Pliobond adhesive while the other end is looped over the top of the dial face. This provides a pointer.

alignment

Alignment procedures for vhf equipment are fairly standard and should be familiar enough not to warrant a detailed discussion here. However, there are a few points that it might be well to underscore.

I-f: Apply a modulated 10.7-MHz signal to the input to the i-f amplifier through a 10-pF capacitor and adjust the i-f transformers for proper waveshape at the input to the detector transistor. (Use a scope capable of displaying 10.7 MHz.) If a scope is not available adjust the i-f for maximum audio at the speaker. Keep reducing the input signal to the minimum discernable level when peaking. Connect the converter output to the i-f amplifier input.

Converter: Apply an accurate, modulated 144-MHz signal source to the antenna jack. Set the compression trimmers about midway through their range. Set the dial at the point where you want 144 MHz to be and adjust the 50k pot on the bottom of the voltage divider string to about 4.5 volts. Squeeze or spread the turns on L4 until the 144-MHz signal is heard. (It may be necessary to adjust the 50k pot again slightly.) Peak transformer T1; peak L1, L2 and L3 in the same manner as L4.

Move the dial to where 148 MHz will be. Set the signal generator to 148 MHz and adjust the trimmer across L4 so that the 148 MHz signal can be heard. The tuning voltage should be close to 9 volts. Adjust the trimmers across L1, L2 and L3. Since these adjustments interact you will have to go back and readjust the 144-MHz setting.

Repetition of the above steps will gradually reduce the amount of adjustment necessary each time until the set is aligned. Remember that the inductance sets the lower end and the trimmer capacitors set the high end of the tuning range. As with the i-f strip, keep the input from the signal generator as low as possible to prevent overdriving.

Transmitter: The important thing to keep in mind when aligning any solid-state transmitter is to apply reduced power initially. In this case use about 9 volts. Connect a wattmeter to the output of the amplifier and, starting with the oscillator, peak each stage for maximum. It might be necessary to back L5's slug off slightly from its peak setting to insure oscillator starting every time power is applied.

Throughout tuneup procedure momentarily remove the crystal to insure that there is no self-oscillation. Gradually increase the power supply to full voltage. With a dummy load on the antenna, a nearby receiver can be used to set the 5k modulation control. With normal voice modulation increase the pot until splatter is noted on the receiver, then backoff slightly. An actual on-the-air check will provide the basis for further adjustments.

references

1. Henry Cross, W100P, "An Obsolete 50-MHz Mobile Receiver," *QST*, December, 1967, page 31.
2. Doug DeMaw, W1CER, "The Connecticut Bond Box," *QST*, August, 1968, page 11.
3. Donald Nelson, WB2EGZ, "High-Performance 2-Meter Converter — The Two-Meter Winner," *ham radio*, August, 1968, page 22.
4. Richard Preiss, W7HCV, "The Two-Meter QRP Mountain Topper," *QST*, May 1970, page 11.
5. Robert Thompson, K1AOB, "From Breadboard to Printed Circuit — The Easy Way," *73*, February, 1970, page 36.

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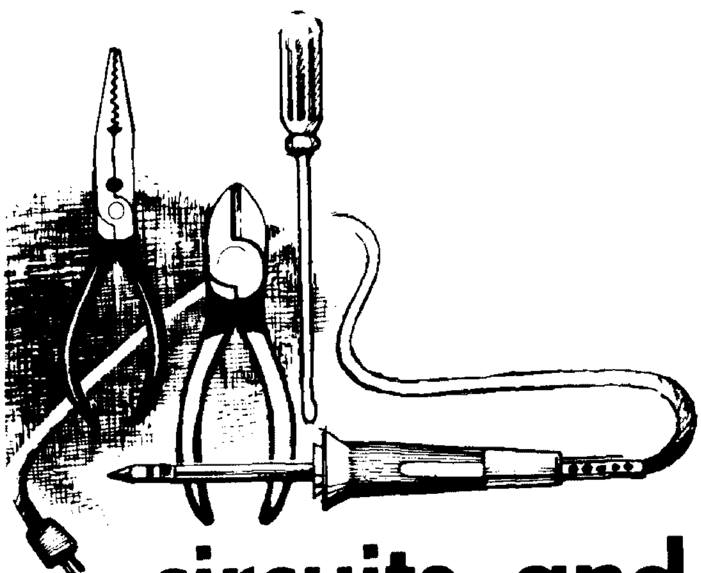


circuits and techniques

ed noll, W3FQJ

vhf beacons

The RSGB and various European groups and clubs are to be congratulated on their vhf beacon facilities. Even little Iceland has a beacon station. In the study of vhf propagation and early forecasting of openings they are marvelous, and they permit a high order of systematic and recorded scientific investigation of sporadic-E, abnormal tropospheric, aurora and other propagation modes. It might well be that they are putting the English and European hams well in the forefront in these endeavors.

The value of the vhf beacons is not confined to propagation studies *alone*. These signals are transmitted continuously and are always available for amateur tests and measurements. They are a great benefit in checking antennas and receiver front ends and as frequency check points. Also, it would not be an unsurmountable problem to adapt them for use as satellite test-signal sources when the era of long-distance vhf communications begins for the radio amateur.

There is abundant unused spectrum in our vhf/uhf bands. The Technician license was initiated in hopes of some concerted vhf-uhf experimentation. A system of beacons could be a splendid scientific tool.

The caliber of the RSGB effort is disclosed in an article by R. A. Ham in the June 21, 1971 issue of the *British Electronics Weekly.*[1] For example, in the study of tropospheric changes, a barograph chart of beacon signal level vs barometric pressure is kept. Mr. Ham's tests indicate that when the pressure rises to the 30-inch level, and then begins an additional steady rise, a likely opening can be anticipated at the time the pressure begins to fall. In fact, there is some rise in the magnitude of local signal in the latter period of the rise and then long-distance openings are set off with the moment of decline.

In other checks, Mr. Ham was able to monitor European, English and Icelandic beacons continuously under sporadic-E conditions. He was able to come up with various patterns of range and direction changes that resulted from the E activity.

phase-locked operation on vhf

The vhf potentials of the phase-locked loop were mentioned in a previous

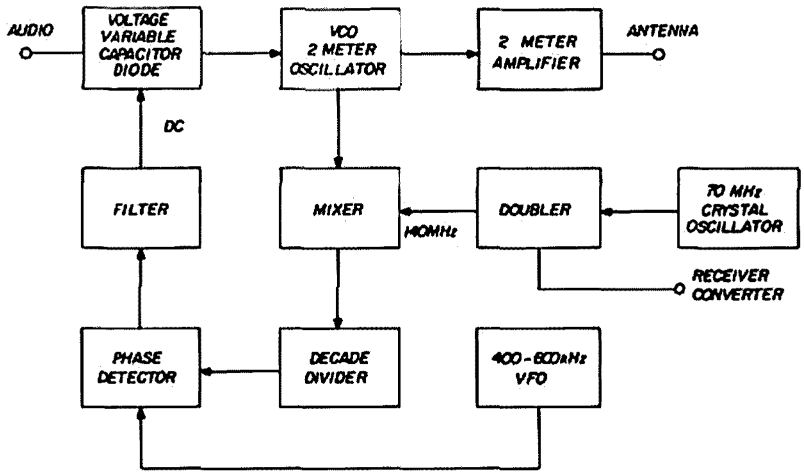


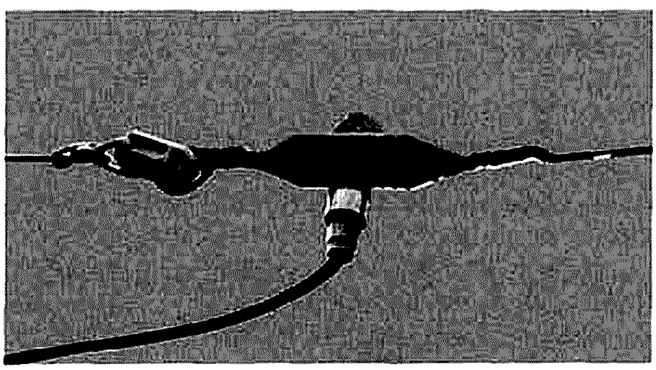
fig. 1. Two-meter phase-lock facility used by G3WXO.

column. G3WXO[2] uses a decade counter (Motorola SN7490N) and a quadruple two-input positive NAND gate (Motorola SN7400N) connected as a phase detector in the lock path (fig. 1). No multiplier chain is used; the fixed frequency is generated directly by a voltage-controlled two-meter oscillator. The oscillator output supplies signal to a follow-up two-meter amplifier.

A portion of the output is applied to a mixer along with a 140-MHz crystal-controlled component. The difference frequency (4-to 6-MHz range) is applied to the decade counter.

The phase-detector compares the output of the counter with the output of the low-frequency variable-frequency oscillator which is tunable from 400 to 600 kHz. This oscillator is the vfo control for the transmitter. The dc-control voltage at the output of the phase detector is applied to the voltage-controlled 2-meter oscillator by way of an appropriate filter.

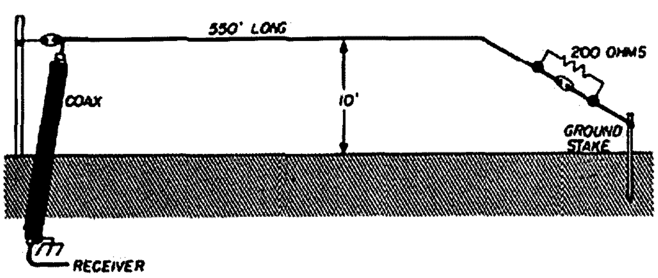
A varactor-diode circuit responds to the dc-control voltage. It is possible to change the frequency of the voltage-controlled oscillator between 144 and 146 MHz by varying the frequency of the vfo. Furthermore, an audio signal can be used to frequency modulate the 2-meter oscillator directly. There is no trouble in obtaining full deviation with the center frequency held fast by the phase-lock system. In G3WXO's two-meter station the 140-MHz component is also used as local oscillator injection for his two-meter converter.


Feed end of the Beverage antenna.

beverage antenna

I covered the subject of anti-QRM receiving antennas in a previous column.[3] One of the antenna types mentioned was the Beverage. Its possibilities for receiving from a single general direction with mini-

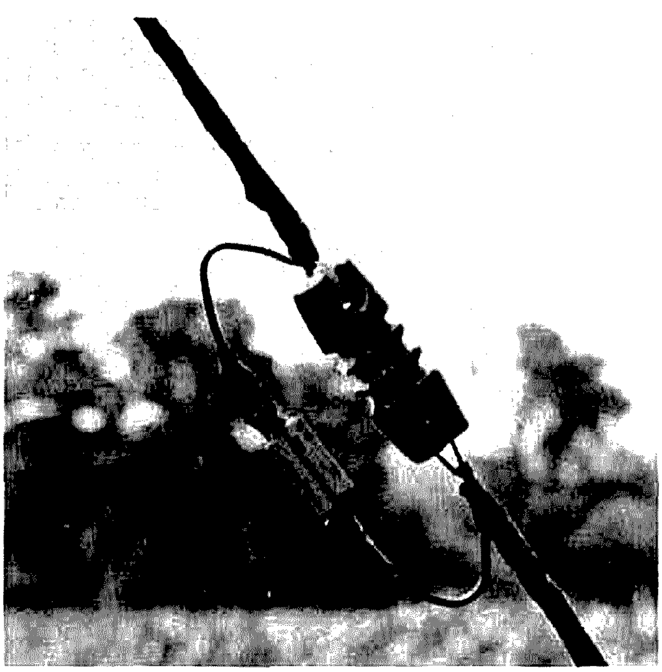
fig. 2. Test Beverage antenna at W3FQJ.



mum pickup from side and back were verified with a Beverage antenna at W3FQJ.

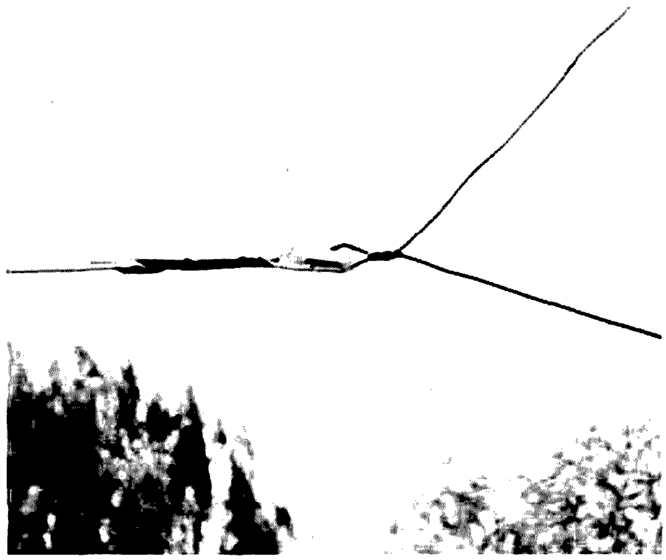
The antenna was short as Beverage types go; total length was 550 feet, fig. 2. Height at no point was more than 10-feet above ground. The end was terminated in a 200-ohm non-inductive resistance (three 600-ohm resistors in parallel). A ground stake was driven 4 feet into the ground. Approximately 120 feet of coaxial cable linked the feed end of the Beverage to the shack. The inner conductor of the coaxial cable was connected to the Beverage wire; no connection was made to the outer conductor at the antenna end. A coaxial switch permitted the receiver input to be switched rapidly between the Beverage antenna and the combination transmit-receive antennas used on the various amateur bands. Here are the results:

1. The Beverage antenna was highly directive off the far end. It favored low-angle long-distance signals over short-skip and local signals.



Terminated end of the Beverage antenna.

2. In general the received signal strength was lower than a good operating antenna on any specific band. However, back and especially side pickup was in most cases, less than similar pickup with the comparison antenna. That is, the ratio of the signal from desired direction over interference from other directions was higher than with the comparison antenna.



Construction used at the base angle of each element in the triangle beam.

3. Similar results were obtained on all bands 10 through 160 meters. The antenna was pointed about 15° south of west. Maximum direction therefore was directly across the continental United States and showed a definite preference for W5s, WØs and southern W6s.

Some directivity existed even on 160 meters despite the fact that the antenna should not be considered a true Beverage for this band since it is only about one wavelength long. Nevertheless, it provides a good cutback of W1 signal levels when tuning in signals from the corn-belt states. Vhf checks have yet to be made.

The Beverage antenna can also be loaded for transmitting. The best method of matching is to use the common T-network often used to match single-wire antennas to a transmitter. This network consists of two tapped coils and a single

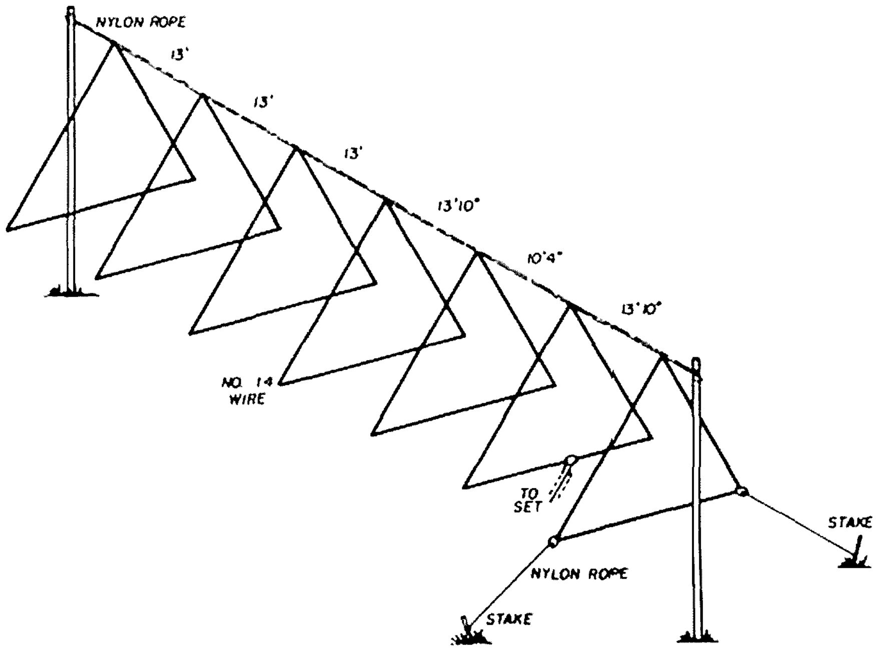
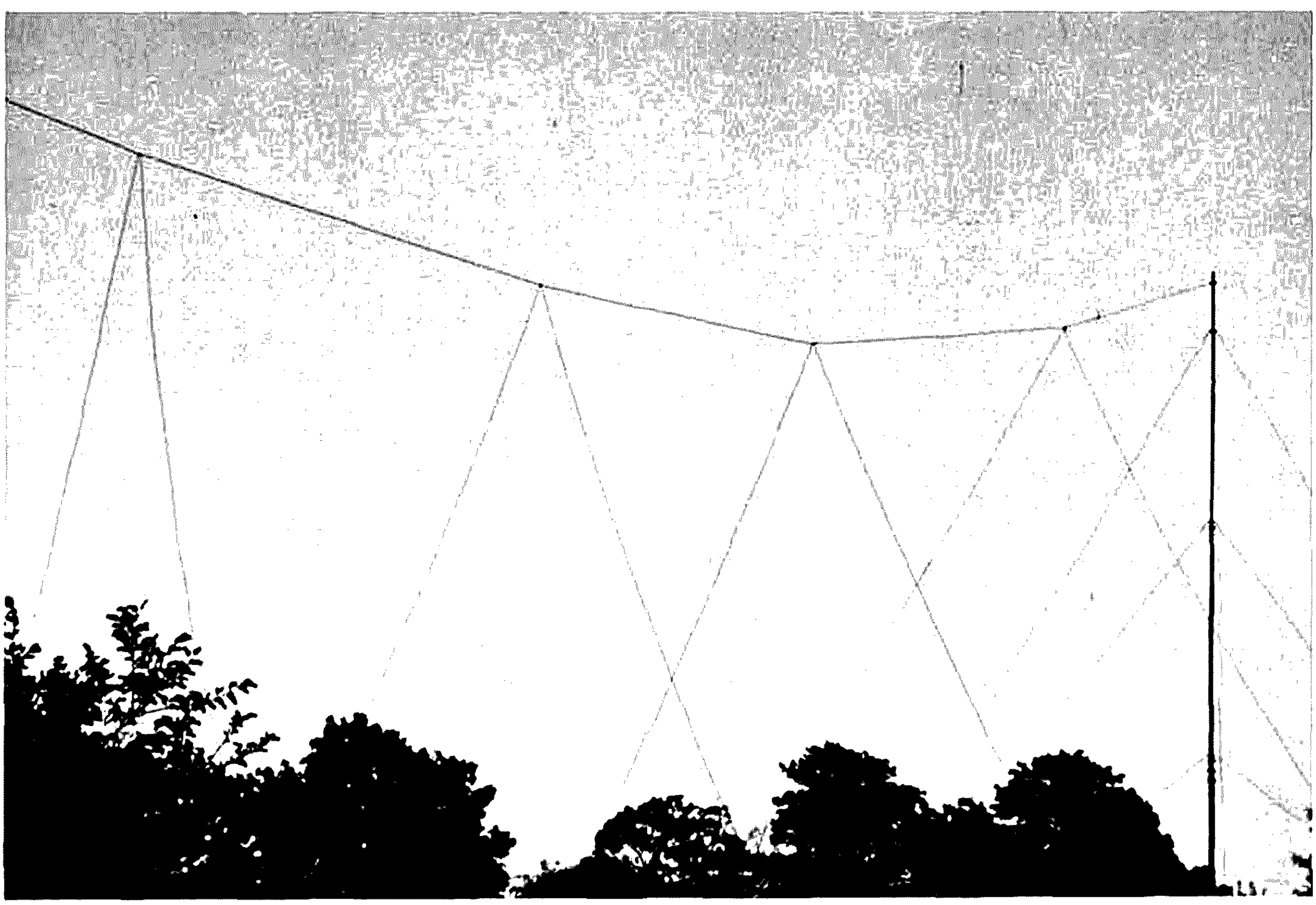


fig. 3. Seven-element 20-meter triangle beam. All elements are staked out the same as the reflector.

variable capacitor adjusted for a good match on each band. There was no trouble with loose rf in the shack with the use of coax between the transmitter and the fed end of the Beverage antenna.

Stations could be raised on each band in the favored direction despite the very low antenna. Reports would be a half S-unit to several S-units below comparison antennas. It was very difficult to raise

Four triangles of the seven-element triangle beam.



anyone off the favored angle. This latter condition checks out the reciprocity theorem, indicating very low sensitivity off the favored direction.

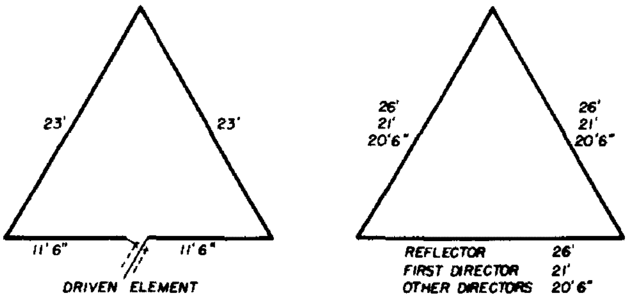


fig. 4. Dimensions of the elements used in the W3FQJ triangle 20-meter beam.

There is substantial work to be done on this style of antenna. The simple tests above just scratch the surface. Other amateurs who have done work with the Beverage indicate that the antenna should not be more than 10 feet above the ground or it will lose its low angle directivity and increase its side pickup.

There is a host of experiments to be done with various types of elaborate ground systems and the selection of length and height above ground to favor a specific band. Added height will increase signal pickup and transmit signal reports. However, just what is that optimum height which *will not permit a serious* cancellation of its anti-QRM characteristics when receiving?

another triangle

A high-gain beam antenna can be built at low cost with the *full-wave* triangle. A 20-meter seven-element triangle (driven, reflector and five directors) was constructed and suspended on nylon rope between two 40-foot telescoping TV masts, fig. 3. All of the closed full-wave configurations (quad, *delta* and triangle) perform well at low antenna heights. This is not to say that they cannot be improved upon by mounting them higher. Of the three types, however, the single-point top of the triangle is a definite advantage for low-cost storm-resistant mounting arrangements.

The seven elements were made of insulated wire. With the apex of the triangle at 40 feet off the ground the feed point of the driven triangle is readily accessible from ground level.

This antenna was positioned to favor Europe and performs well with 200 watts on a band crowded with high beam antennas and kilowatt transmitters. Raising the apex *another* 30 *feet* would permit some spectacular results with QRP power.

more phase-locked loop

Additional checks were made on the phase-locked loop circuit covered in September's "Circuits and Techniques."[4] A dc voltage-control circuit was added (fig. 5) to provide a means of changing the frequency of the voltage-controlled oscillator. Potentiometers replace the variable *capacitors of the* previous circuit. A fixed 250-pF capacitor between terminals 2 and 3 permitted tuning the

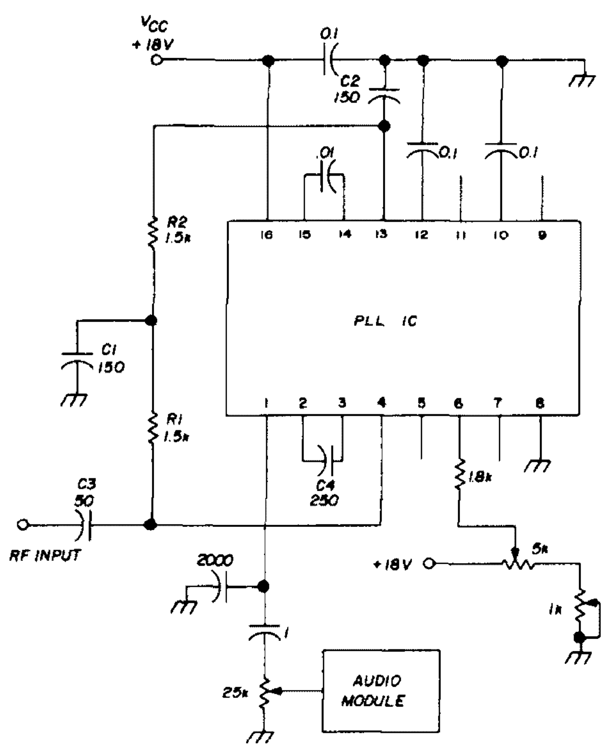


fig. 5. Phase-locked loop receiver with dc vco control.

160-meter band and the high-frequency end of the broadcast band.

Two potentiometers are connected in series as coarse and fine (bandspread)

tuning controls. Excellent bandspreading can be obtained with this arrangement. For example, by using a bandspread potentiometer of several-hundred ohms, a 50-kHz segment of the 160-meter band can be spread over the entire dial. Hand-capacitance effect is virtually eliminated.

A simple fet amplifier improves the sensitivity of the receiver. Strong locals are a problem and sometimes must be trapped-out if you live in an area such as mine with a cluster of broadcast towers not too many miles away.

In one experiment, **fig. 6**, the resonant drain circuit was a replacement superhet rf transformer (such as Stancor 8736). The antenna winding was connected to the input of the phase-locked loop. The

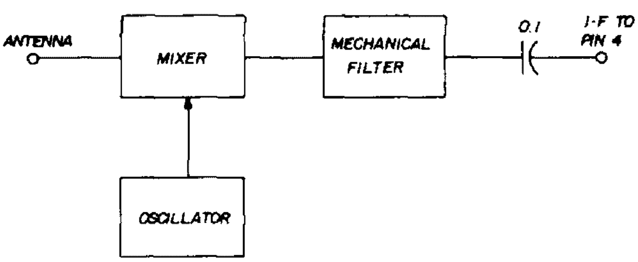


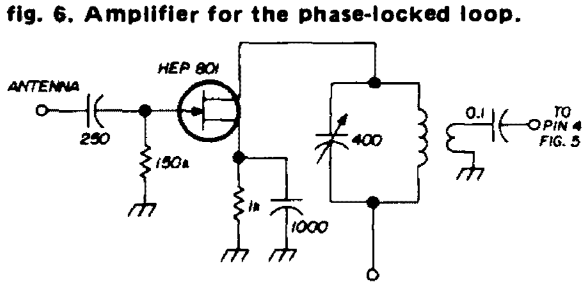
fig. 7. Use of a phase-locked loop in a superheterodyne receiver.

secondary was connected to the drain circuit of the fet. Proper adjustment of the transformer slug permitted operation on 160 meters and a large portion of the broadcast band as well.

Performance was excellent considering the simplicity of the receiver. Sensitivity was excellent with a suitable antenna; selectivity was quite good.

Selectivity is improved with the use of a mixer-oscillator ahead of the device using the phase-locked loop as an i-f amplifier and demodulator. In this arrangement a mechanical filter is placed between the mixer and the phase-locked loop, **fig. 7**.

fig. 6. Amplifier for the phase-locked loop.



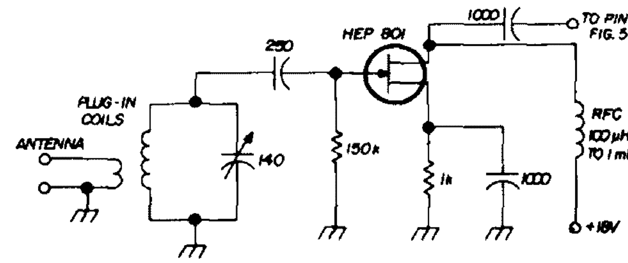


fig. 8. Multiband circuit arrangement for the phase-locked-loop receiver shown in fig. 5.

The arrangement of **fig. 8** was also checked out and gave good a-m performance on the 40-, 80- and 160-meter bands. A radio-frequency choke was connected in the drain circuit, while the resonant transformer was moved to the gate circuit. With this circuit arrangement a tapped resonant coil or plug-in coils can be used for the three bands.

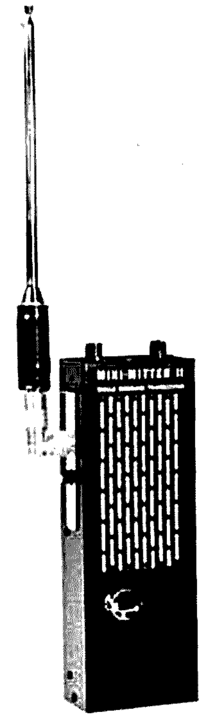
By connecting a trimmer capacitor between terminals 2 and 3 of the PLL, a particular amateur or shortwave band can be located more readily. In this case one potentiometer is used to tune over the entire band and the second potentiometer is used for bandspread tuning within the band.

An amplifier is a definite aid in improving the sensitivity for this simple direct-detection receiver. It is wise not to use an amplifier that can go into self-oscillation; be certain that any amplifier with tuned input and output circuits is properly neutralized before it is connected to the input of the phase-locked loop.

references

1. R. A. Ham, "Radio Society Members Provide Vital Link in Probing Atmosphere," *Electronics Weekly* (England), June 2, 1971, page 14.
2. Jack Hum, G5UM, "Four Meters and Down, from G3WXO," *Radio Communications* (England), June, 1971, page 403.
3. Edward M. Noll, W3FQJ, "QRM-Reducing Receiving Antennas," *ham radio*, May, 1971, page 50.
4. Edward M. Noll, W3FQJ, "IC Phase-Locked Loops," *ham radio*, September, 1971, page 54.

ham radio



Mini-Mitter II

This miniature ssb transceiver for 40 and 80 puts the fun back into amateur radio

Ken Pierce, W6SLQ, 886 San Rafael Avenue, Mountain View, California 94040

Another page in the log book. Time, 2:13. Just moved off the net, traffic complete. Shall I go back to the net or just shut down; A kilowatt just to work 200 miles; what a waste of power. Someone calling me? Another kW? Must be from the sound of the signal. What did he say? Walkie talkie – on 40 meters? I don't believe it.

Those were the musings in my mind that Saturday afternoon I first met Al Clark, W6IHY. A few minutes later I was at the door of American States Electronics, a small neat building in Mountain View, California.

Al, a very affable fellow, showed me around the plant. With justifiable pride he handed me the Mini-Mitter II. This little handful is a complete ssb transceiver that weighs in at 3 pounds, complete with self-contained battery power supply. Unbelievable. I had heard the signal strength and quality of the signal, so I had to believe it.

Al gave me a chance to put it on the air. The signal strength and quality reports were excellent. I worked stations from 300-to 600-miles away with the same consistent reports. I had to have Mini-Mitter II for my own – what did I have to do to get one?

The answer to that question proved to be interesting, and to me, delightful. This transceiver is available in kit form for $149.95, complete. The accessory whip antenna is $7.95. I wondered if I could build the unit and make it work as well as the one I had in my hand. I looked over the kit and wiring instructions and came to the conclusion that it was perfect for me.

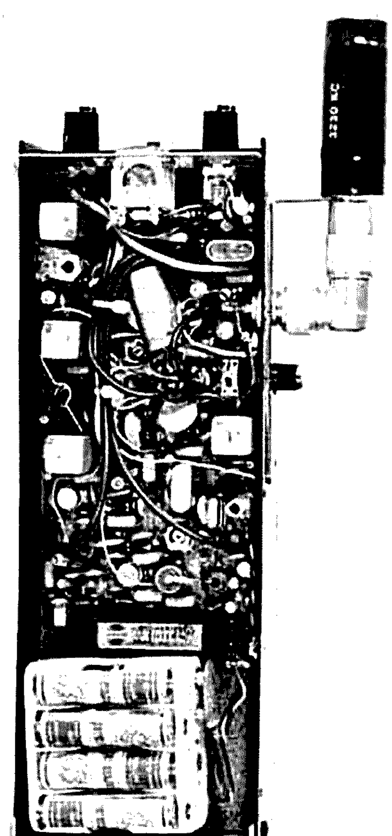
Do not be mislead, the unit contains a high-density printed-circuited board,

about the size of your hand, that contains all the pertinent parts. In the next few hours I was to learn what Al meant by the term, "high-density board."

On the way home, with the kit under my arm, many second thoughts were running though my head. Could I put it together? Would it work? Was the set of instructions as good as I thought? Then I got home and opened the box (giving second priority to my usual Saturday chores), "I" started putting the transceiver together.

I followed the instructions to the letter, no playing it by ear. When I finished, I followed the simple alignment instructions and I was on the air. On-the-air contacts confirmed that it is really easy to have a good quality ssb signal.

The Mini-Mitter II printed-circuit board contains three ICs, 15 transistors and 15 silicon diodes. A Collins mechanical filter with crystal-controlled carrier frequency places the carrier at the proper place on the slope of the filter. This gives



They call it a "high parts density printed-circuit board," but construction is not difficult.

mini-mitter II

frequency	crystal controlled on any 40- or 80-meter frequency
mode	upper or lower sideband
rf output	4 watts PEP
sideband suppression	40 dB or more
output	50 ohms nominal
sensitivity	less than 1 uV for 10-dB signal-to-noise ratio
selectivity	2.1 kHz at 6 dB down
audio output	0.2 watts nominal
power supply	10 volts (eight AA penlight cells, self contained)
size	9 7/8 x 3 3/8 x 1 3/4 inches
weight	44 ounces
price	$149.95

American States Electronics, 1074 Wentworth Street, Mountain View, California 94040.

an audio passband of 350 to 2450 Hz, results in beautiful quality and retains the "presence" of the human voice. The tiny transceiver has many of the refinements of larger, more expensive rigs, including alc, agc, image rejection of 50 dB, 40 dB minimum carrier suppression and 50- to 75-ohm output.

With the whip antenna in place I wondered if I could work anyone just walking around the street. I worked stations as far away as Oakland and San Jose (about 30 miles). One mobile operator had to drive over to see the miniature transceiver for himself.

With such good results with the small whip, I wondered what I could expect from the larger portable antenna — the ASE 17 DL, priced at $66.95. Five states were logged with the portable antenna on its 20-foot mast and I never miss the opportunity to show my logbook to the kilowatt advocates.

I put the little Mini-Mitter II under my arm and went over to see Al again at the factory. I got the same warm welcome I received on my first visit. Al checked out the unit for me, measured the rf output, noted all those things that fascinate engineers, then gave it back to me. We were both satisfied with its performance. I am very proud of this little gem and use it extensively on 40-meter ssb.

ham radio

the ham notebook

electronic keyer notes

The electronic keyer described by VE7BFK in *ham radio*, November, 1969, is superb, and I highly recommend it to home builders. With a dual-lever key the dot memory and dash over-ride features make code operation a pleasure while the construction of the unit is a fine introduction to digital microcircuit techniques. The following notes describe some minor modifications to the original design that have proved worthwhile.

set ciruit

My keyer worked quite well as designed except for an annoying tendency to spawn an occasional extra dot. This problem was solved by reducing R14 from 4.7k to 1.2k ohms; this gives a set pulse of only 200 microseconds which is quite sufficient for reliable operation of the dot memory chain even with the supply line reduced to 2 volts.

keying circuit

The original keyer is all solid state and uses a switching transistor to key the transmitter. When grid-block keying is used, as is generally the case, the keyer ground is at a different potential than the transmitter ground. Apart from presenting a minor safety hazard I found that this practice made it difficult to keep rf energy from triggering the keyer. For this reason I substituted the reed relay and driver circuit shown in **fig. 1**.

Operate and release times are less than one millisecond, giving negligible shaping at any practicable keying speed. The resistor in the keying lead limits current pulses and prevents sticking contacts. I wound my own bobbin around a miniature reed capsule and ended up with a coil resistance of 60 ohms. This operated nicely with a series resistor of 82 ohms to recude the drain on the power supply; higher resistance relays should not need this resistor. The diode across the coil damps the spikes of back emf which might otherwise feed back into the circuitry and play hob with the timing.

monitor

The circuit as described had no monitor. **Fig. 2** shows the monitor in use in

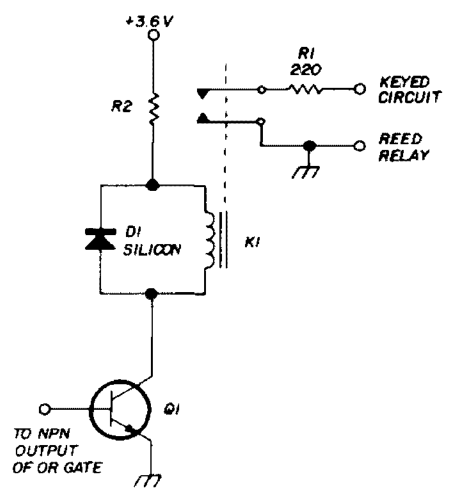


fig. 1. Relay driver circuit. Q1 may be any small-signal silicon type such as the 2N3643 or BC108 with 300 mW or greater rating. R2 is current-limiting resistor, not required if relay K1 has coil resistance greater than 100 ohms.

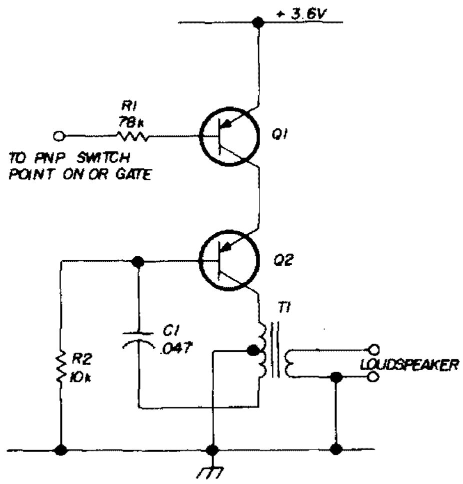


fig. 2. Monitor circuit. Transformer T1 is a transistor output transformer, 500 ohms center-tapped to 16 ohms. Q1 and Q2 are pnp small-signal silicon transistors such as the 2N3638 or BC118.

my keyer. It is a useful addition even if your transmitter has a built-in sidetone oscillator as it provides an audible check on keyer operation when setting it up. It also permits you to learn the squeeze-keying technique *before* you try it on the air. Transistor Q1 functions simply as a switch to key Q2 which is a form of relaxation oscillator. It may be necessary to juggle the values of R1 and R2 to get the best sounding note. The transformer is not critical and both the output and driver transformer from a junked transistor radio have been used with success.

components

Timing capacitor C3 in the original clock circuit is shown as a 1-μF electrolytic. This can be a source of trouble, and I recommend the use of a polyester 1-μF type at this point. A more suitable value for the speed control R7 will then be 100k ohms; R6 may also need to be increased, depending on the speed range that your own particular operation demands.

I used the Motorola RTL MC790P dual flip-flop and two MC724P quad gates as suggested by VE7BFK. These have worked very well, but experimenters may like to try the new TTL devices manufactured by Motorola and others. These devices claim a superior specification at about half the cost of the already reasonably priced RTL devices. If you want to run the keyer from a battery supply follow VE7BFK's advice and use the special milliwatt ICs with the original solid-state keying.

If you don't want to use the printed circuit board specified in the original article I would suggest the use of Vector board made especially for dual-in-line packs. The board has the power and ground rails already set up, and wiring is much less of a chore than it is with standard Vector board.

Barry Kirkwood, ZL1BN

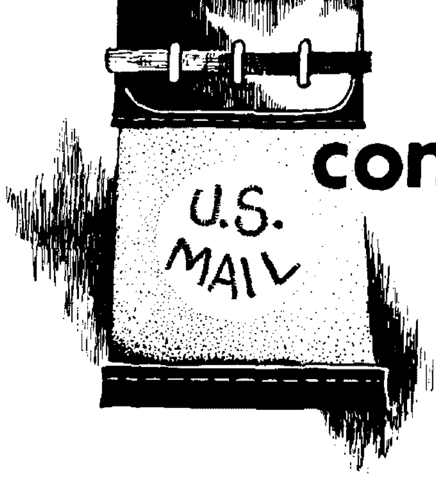
heath ten-minute timer

The Heathkit SB-630 Station Console incorporates a timer which uses a raucous 6.3-volt buzzer, Heath number 69-38. This buzzer has some very fine wire connections to its terminals, which appear to break off as a result of the vibration of the buzzer, mounted on a rubber grommet. Disassembly of the buzzer proved to be impossible without damage, and the unit was not sufficiently attractive to replace with an identical unit.

After some *value engineering* the matter was handled by putting a resistor from the 6.3-volt buzzer lead to the loudspeaker phono jack used on the phone-patch portion of the unit. This resistor can be from 30 to 100 ohms, depending upon the power capabilities of the loudspeaker, and the amount of sound desired. A junkbox resistor, therefore, proved to be more satisfactory than spending money to replace the damaged buzzer.

Incidentally, the instruction manual keeps mentioning a ten-minute interval for reminding the operator to identify. Unfortunately, this will lead to violation of the FCC regulation. The timer should be set for some shorter time so that the identification will be made before 10 minutes have expired, even if the other station is transmitting at the end of the ten minutes.

Bill Conklin, K6KA



comments

smith charts

Dear HR:

I was delighted to see your fine article on the use of the Smith chart in the November, 1970 issue. Your generous sprinkling of the step-by-step instructions with full-fledged graphs to illustrate each use should go a long way toward inviting many amateurs to learn how to use this powerful tool. Perhaps your article will become the implement for helping them come to grips with the complex Z, and to discover that Z often has a jX as well as R.

My daily work involves approximately equal use of the Smith Chart and the slide rule, and as a result I am equally at home with many phases of its use. I can therefore appreciate the opportunity you are presenting to those who might otherwise never discover that all those circles aren't really all that difficult.

Without intending to detract in any way from the otherwise excellent presentation, I would, nevertheless, like to call your attention to two basic errors. One is in the matching-stub example and the other in handling attenuation through a lossy line.

In the matching-stub problem, the procedure is in error because impedance operation is used, rather than admittance. Stub use implies *shunt* connection to the line, and indeed, the schematic diagram shows the shunt connection. Thus there can be no disagreeing that admittance procedure is required to obtain a correct answer to the problem as presented.

To illustrate the required changes, refer to the Smith chart in fig. 1 above. The proper steps are as follows:

1. Normalize the load impedance
   $Z_L' = (32 + j20/)/50 = 0.64 + j0.4$

2. Locate this point on the chart and draw a line through it and the chart center, extending the line through the peripheral scales in the negative, or bottom, portion at $0.336\lambda (a^\circ = -62^\circ)$.

3. Construct a constant-gamma circle through Z_L, on through the admittance point Y_L, and intersecting the unity *conductance* circle (G = 1) at point A.

4. Draw a line from the chart center through point A to the outer scale at 0.348λ(or $a^\circ = -71^\circ$). L_d, the distance from the load to stub, is the distance from 0.336 to 0.348.
   $L_d = (0.348 - 0.336) = 0.012\lambda$
   $a^\circ = 71^\circ - 62^\circ = 9^\circ$ (4.5 electrical degrees)

5. To find the length of the stub, determine the amount of susceptance necessary to match out the load. The required susceptance is the difference between the susceptance at point A and the susceptance at the center of the chart. The susceptance at point A is $-j0.67$. The required stub susceptance is

$$\beta = +j0.67$$

6. Determine the equivalent stub reactance by taking the reciprocal of the susceptance (as described in *example 4,* page 21, November, 1970)

$$X = -j1.49$$

7. Locate the reactance $-j1.49$ on the rim of the chart (point B). Determine the distance between the short-circuit point and the required reactance (point B) along the "wavelengths toward generator" scale. $L_s = 0.344\lambda$. ($a^\circ = 248^\circ$, 124 electrical degrees).

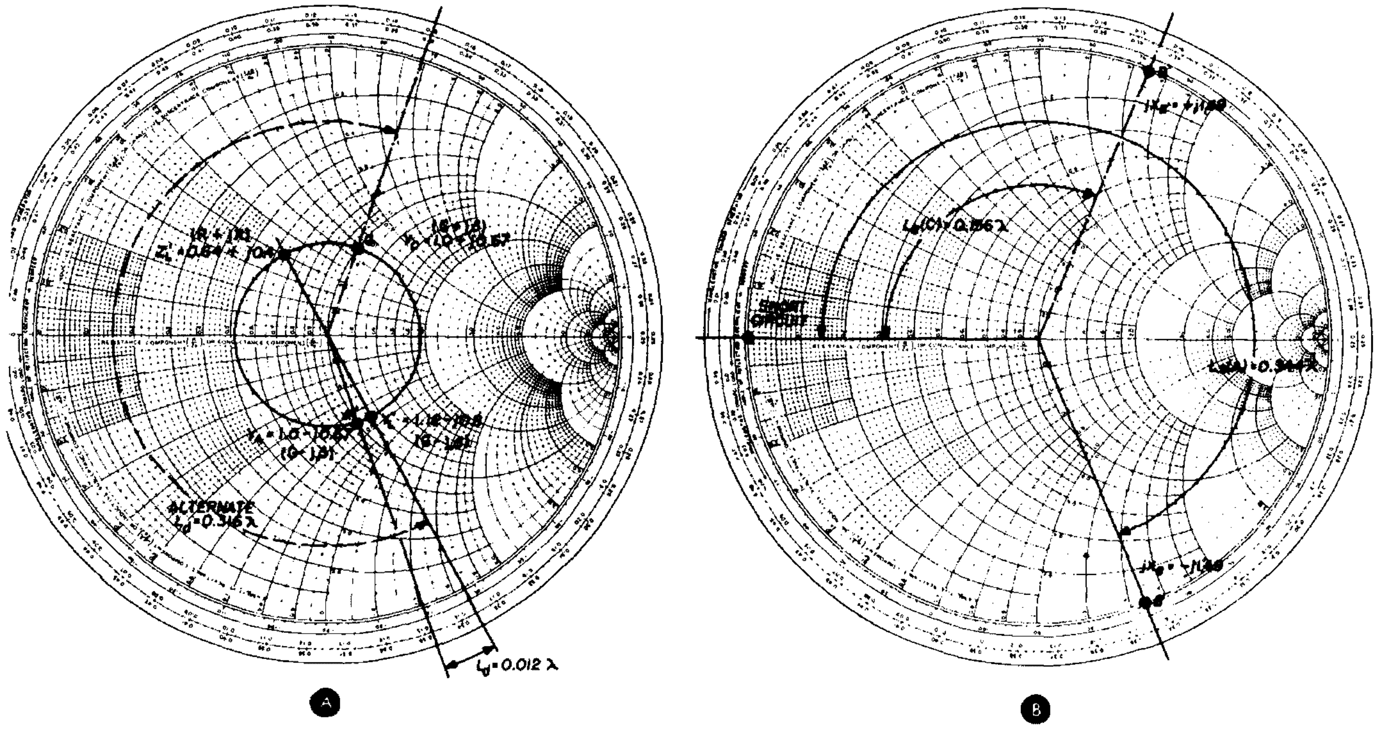
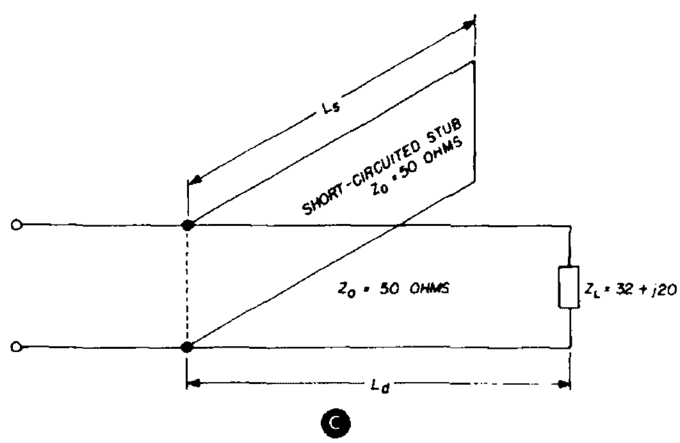


fig. 1. Using the Smith chart to find matching stub length and location.



For practical reasons it may not be possible to place a shunt stub only 4.5° from the load. It may be necessary to increase the distance L_D to the next point where G = 1 (not R = 1), represented by point C, fig. 1. In this case L_D would be measured, clockwise from 0.336 through 0.50 to 0.151. Using the reflection coefficient scale, from a° = −62° to 180° plus 180° to +71°, which totals 227°R, or 113.5 electrical degrees. This represents 0.316λ. This will require a +jX stub, length shown as L_s(C), of the same numerical reactance value as before.

lossy lines

Turning now to the lossy-line problem in which the line attenuation is 2.0 dB, the error concerns the method used in determining the corrected voltage coefficient for the lossy condition. You have stated incorrectly in both **Steps 4** and **5** that the voltage reflection coefficient in the lossy case is 2.0 dB lower than that of the lossless case. This is incorrect because the voltage reflection coefficient varies directly with *power ratio* of one-way line attenuation and *not* the voltage ratio.

Your ρ* correction factor of 0.794, derived from

$$2.0 \text{ dB} = -20 \log_{10} \frac{E_{output}}{E_{input}}$$

yields an input reflection coefficient for line attenuation of exactly 1.0 dB; *not* 2.0 dB. The correct correction factor is 0.631, derived from

$$2.0 \text{ dB} = -10 \log_{10} \frac{\text{Power out}}{\text{Power in}}$$

This is evident because by using the radially-scaled parameter titled "Transmission Loss, 1-dB steps," the interval

*ASA standards have adopted ρ (rho) to represent the magnitude of the reflection coefficient.

shown on your plot (to represent 2.0 dB) measures 1.0 dB on the scale. Plotting off an actual 2.0 dB with the 1 dB-step scale brings the Z_L (lossy) to 1.08 + j1.05, which indeed has a reflection coefficient of 0.631.

According to Phillip Smith, inventor of the Smith chart, "Since the propagation path is common to both the forward and reflected wave energy, the latter, in its backward path to the initial point of entry, will be attenuated in the same ratio as was the incident wave energy. At the initial point of entry the power reflection coefficient is, thus, a measure of the two-way transmission loss, expressed as a power ratio. One-half of this, therefore, represents the one-way transmission loss, viz.,

$$dB = \tfrac{1}{2}(-10 \log_{10} \rho^2) = -10 \log_{10} \rho \ldots.$$

The line in example 8 has a one-way attenuation of 2.0 dB. That quantity of power which will ultimately be reflected back to the input suffers a 2.0 dB loss in the forward direction and another 2.0 dB during the return trip. The power returning to the input (ρ^2) is therefore 4.0 dB below its original level at the beginning of the journey. Thus ρ^2 is 4.0 dB below what it would be with a lossless line, and so also would be ρ. Therefore, to derive the ρ correction factor, we may write (all logs to base 10):

$$\begin{aligned} -10 \log \rho^2 &= 4.0\text{ dB} \\ -\log \rho^2 &= 0.4 \\ \rho^2 &= 0.398 \\ \rho^2 &= \sqrt{0.398} = 0.631 \end{aligned}$$

But a *one-way* loss must be, as Smith has stated:

$$\begin{aligned} \tfrac{1}{2}(-10 \log \rho^2) &= 2.0\text{ dB} \\ -\tfrac{1}{2} \log \rho^2 &= 0.2 \\ -\log \rho &= 0.2 \\ \rho &= 0.631\text{, and not } 0.794 \end{aligned}$$

However, $0.794^2 = 0.631$; it is evident that the corrected input reflection coefficient for the line with 2.0-dB attenuation terminated with a load whose reflection coefficient is 0.68 will be the product of 0.631 x 0.68 = 0.429, in contrast to the result of 0.54 as indicated in the article.

Walter Maxwell, W2DU
Dayton, New Jersey

triangle antennas

Dear HR:

The triangle antenna is a good antenna. However, I always wonder why loop-antenna articles never include the fact that a full-wave loop has 2-dB gain over a dipole ("Quads and Yagis," *QST*, May, 1968). A square quad or triangle delta configuration probably won't provide all of the 2-dB gain, but then cost and mechanical considerations generally take precedence over theory.

Wayne W. Cooper, K4ZZV
Miami Shores, Florida

teacher exchange

Dear HR:

I teach electronic, electrical, radio and tv, and mathematics subjects at various levels at the County Technical College in Norfolk, and I am seeking a one-year teacher exchange with an American teacher. I have been selected on this side but so far we have been unable to find an American teacher (and family) desirous of working and living in England (but retaining U. S. salary) for a year beginning August, 1972.

If any of your teacher readers (perhaps, but not necessarily a radio amateur) would be interested in an inexpensive one-year "holiday" in England, I invite them to write to me or contact my "exchange manager," WB2FBF, who will answer any local queries. Official details and application forms in the U. S. are available from the Office of Education, Washington, D. C. 20202.

David Lake, G3ZCA
County Technical College
Tennyson Avenue, King's Lynn
Norfolk, England

coax-cable leakage

Dear HR:

I read with interest VE7ABK's article in the March, 1971 issue. His comment about shielding (or lack of it) in RG-8/U brought to mind a fact that seems to be little known among amateurs. Common single-braid coaxial cables such as RG-8/U leak like a sieve.

Many times this leakage does not matter, or the effect is not noticed; however, in some cases, as with repeaters or impedance bridges where the oscillator and detector levels are more than 150 dB apart, lack of shielding in ordinary coaxial cables and connectors is of major importance. Double-braid cables such as the old RG-9/U and RG-55/U, and the newer versions, RG-214/U and RG-223/U, are made for just such circumstances. Also, type-N connectors are much better than BNC.

Ron Guentzler, W8BBB
Ada, Ohio

rf speech clipper

Dear HR:

There are two errors in the schematic diagram of the rf clipper described in the August, 1971 issue (page 19). The fixed capacitor across the rf input to the gate of the 2N5248 source-follower stage should be 100 pF instead of the 0.01 μF shown. The 100-pF capacitor pads the tuning range of the Arco 403 trimmer to provide 130 pF, the resonant tuning capacitance for the mechanical filter.

The second error is in the 2N2222 amplifier stage. The emitter of this transistor should be bypassed with a 0.01-μF capacitor to provide sufficient ac gain to allow clipping. It has come to my attention that it may not be possible to develop sufficient gain with the untuned 2N2222 collector. This is due to a combination of lower gain devices and/or the ommission of the source-follower input stage when used in a transmitter with high-frequency crystal filters (such as the Heath SB400 series).

In such cases a 455-kHz resonant circuit with a loaded Q of 10 to 20 is required in place of RFC3. This source follower provides +10 dB power gain (voltage gain slightly less than unity). Without this stage there may be lack of drive to the clipping diodes. For Heath and Kenwood transmitters, an additional power amplifier stage should be added between the 2N2222 amplifiers and the clipping diodes.

One other point: the part number of both RFC1 and RFC2 in fig. 1 should be 3E2A, not 3K2A.

Bruce Clarke, K6JYO
Fullerton, California

ac line cords

Dear HR:

A recent *ham notebook* item from K6JYO described a "safer suicide cord." Many of us are guilty of removing ac power cables from wall outlets by the simple and quick expedient of jerking the cable, rather than walking over to the ac plug. One of the dangers involved, unknown to most of us, is that when pulled from some distance, the cable acts like a whip — and you can LOSE AN EYE! I know of one such case. It might be wise to stop this practice and to remind others as well.

Frank Hatanaka, W6EG

using the mailing wrapper

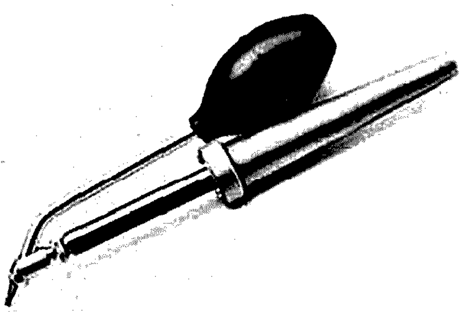
Dear HR:

When my copy of *ham radio* arrives each month I slit the mailing wrapper on the open-page side and leave it attached to the magazine. On the wrapper I write the title and page number of any article that interests me. After much handling, particularly if I build something from the issue, the wrapper is well stained with solder flux and finger marks. When I remove the wrapper I have a nice clean magazine to put with the rest of my *ham radio* file.

Robert J. Farnum, W4PWC
Miami Shores, Florida

new products

desoldering iron



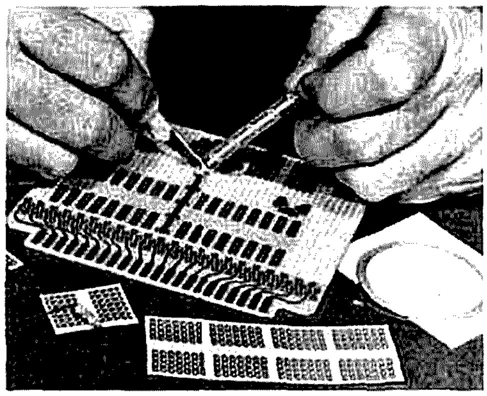
Weller has introduced a new hollow-tipped desoldering iron designated the DS-40. A vacuum bulb removes solder melted by the tip; this aids parts removal from printed-circuit boards or conventional wiring.

This new tool can also be used to solder in new components once the old ones have been removed. The hollow tip fits over wire leads giving 360° contact. This gives more uniform heating and better solder joints.

The unit is designed for professional as well as amateur applications. Replacement tips are available in a variety of sizes and a three-wire version is also available. The standard two-wire version is sold with a tip and vacuum bulb for $14.50.

Use *check-off* on page 120 for more information or write to Weller, 100 Wellco Road, Easton, Pennsylvania 18042.

circuit-stik



A new concept of instant printed circuits now permits you to rapidly build projects directly from a schematic diagram. This new approach to printed circuits consists of a complete family of circuit subelements and associated circuit materials. With instant printed circuits there is no messy etching, and in most instances, no need to drill any holes.

The instant printed-circuit subelements consist of printed conductive patterns on a very thin epoxy-glass board backed with pressure-sensitive adhesive. Subelements are available for all types of integrated circuits, transistors and other components. Any combination of circuit element configurations can be mixed on one board; The circuit subelements are all pre-drilled with holes on a 0.100-inch grid to match the pattern of 0.100-inch Vector board.

As an example, assume you are building a project using a 10-lead, TO-5 integrated circuit. Simply pick up the matching subelement, strip off the protective backing and stick it in position, matching the holes in the printed-circuit subelement with holes in the Vector board.

The printed-circuit subelement has pre-drilled holes to accept the integrated circuit leads and provisions for connecting components or leads to the integrated circuit. From this point, there are two basic ways to complete the circuit; or a third approach that uses a combination of the first two. One way to complete the

circuit is simply to use jumper wires and component leads. Or, you can use conductive tape for making inter-connections and donut pads for terminating component leads.

Just place donut pads at points where you wish to terminate components. The components terminated on the donut pads should be mechanically secure without support from the pad. That is, resistors and capacitors should be pushed through holes in the board so that the component rests firmly on the Vector board.

To use the conductive tape, hold one termination point with an X-acto knife and strip the protective paper as the tape is laid down. The knife blade is then moved to the opposite termination, pressed down, and the tape is popped off with a quick tug. For best adhesion, roll the tape down with the side or heel of the knife or other burnishing tool. The adhesive on the copper tape is electrically conductive, and, provided it is burnished for good adhesion, no soldering is needed for temporary patches and connections. However, to eliminate any possibility of opens or intermittents, a drop of solder is highly recommended at termination points.

Various assortment packets of Circuit-Stik are available from Circuit Specialists Company, Box 3047, Scottsdale, Arizona 85257 at prices of $5.50, $7.95 and $9.95; add 35 cents for shipment by air mail. For more information use *check-off* on page 120.

hep semiconductor catalog

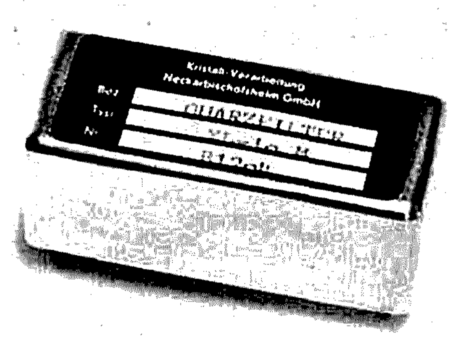
More than 31,000 semiconductor devices are cross-referenced to HEP replacements in the new 1971 Motorola HEP Semiconductor Cross Reference Guide and Catalog. Included in the catalog are 1N, 2N, 3N, JEDEC, manufacturers' regular and special "house" numbers and many international devices, with particular emphasis on Japanese types.

Four hundred and seventy-one HEP

items are included in this guide, including kits, books and accessories. As in previous editions, the Motorola HEP devices are listed by type number with a packaging index, device dimension drawings and selection guide information.

This cross-reference guide and catalog is available free at local HEP suppliers throughout the country. It should be of particular interest to the amateur, hobbyist-experimenter and the professional service dealer since it gives the minimum/maximum ratings and the electrical characteristics for the HEP devices as well as cross-reference information.

no-ring crystal filter



An extremely-sharp ten-pole ladder filter for narrow-band CW and pulse operation has been introduced by Spectrum International. The new model XL 10M is manufactured by KVG of West Germany and eliminates the ringing effect commonly found in some narrow bandwidth crystal filters by a near-Gaussian response to -6 dB. The filter comes with built-in input and output transformers in its hermetically sealed enclosure. The filter features a bandwidth at -6 dB of 500 Hz, symetrical around a center frequency of 9 MHz and with a shape factor (6:60 dB) of 2. It has a maximum insertion loss of 10 dB and a minimum ultimate rejection of 80 dB. It sells for $59.95.

For complete technical specifications write to Spectrum International, Box 87, Topsfield, Massachusetts 01983 or use *check-off* on page 120.

More Details? CHECK—OFF Page 120

audio filters

Kojo audio filters can greatly improve reception on all receivers by removing high-frequency audio hiss, background noise and ssb *buckshot*. The ssb filter uses a low-pass design with sharp cutoff to provide rejection better than 30 dB at all frequencies above 3500 Hz. The ssb filter is specifically designed for placement in the low-impedance line to headphones or speaker.

The Kojo cw filter has a spot frequency of 780 Hz and a passband of 1100 Hz with a reference level 40 dB below the signal level at the design frequency. The peak of the passband is 100 Hz wide at the -3 dB points. The cw filter is designed for low-impedance input and high-impedance output; high-impedance crystal-type earphones are recommended.

The Kojo audio filters use top grade coils and quality components, and are available in kit form or ready-to-use deluxe unit enclosed in a cabinet. The cw filter kit is $7.95 (deluxe cw filter, $15.95); the ssb filter kit is $11.95 (deluxe ssb filter, $19.95). Postpaid from The J. Lynch Company, Post Office Box 7774, Phoenix, Arizona 85011. For more information use *check-off* on page 120.

allied catalog

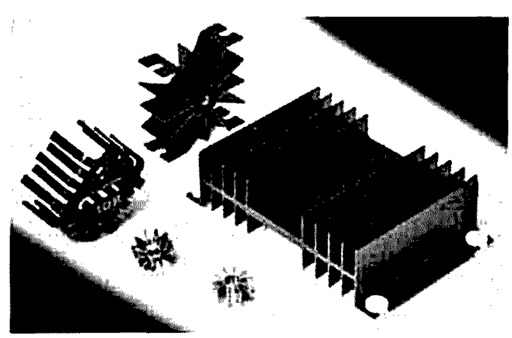
The new 132-page, *1972 Electronic Parts and Accessories Catalog* has just been announced by Allied Radio Shack. The catalog lists thousands of hard-to-find electronic items, accessories and repair components in addition to the complete line of Knight-Kit, Science Fair, Allied, Realistic, Archer, Micronta and Radio Shack brand products.

The new catalog, numbered 215, is designed for the amateur, hobbyist, kit builder and electronics professional. It is available free on request from Allied Radio Shack, 2725 West Seventh Street, Fort Worth, Texas 76107 or by using *check-off* on page 120.

electronics hardware

International Rectifier Corporation has introduced a new line of semiconductor hardware for home experimenters. The line includes integrated-circuit and transistor sockets, universal heat dissipators, heat exchangers, semiconductor mounting kits and a dual-in-line component carrier.

A total of eight sockets are available, including three matched integrated-circuit sockets for 8-, 10- and 12-pin TO-5 case styles. Prices are 65 to 75 cents in assorted quantities. Low profile, dual-in-line IC sockets will be available in 14- and 16-pin sizes at 80 to 99 cents in assorted quantities. This is the industry's first full assortment of sockets for every hobbyist purpose.



The International Rectifier universal transistor sockets are designed to be used with in-line plastic SCRs, triacs and some 19 transistor bases. Sockets for power transistors feature high reliability and brass/cadmium plated contacts, with a PC board drilling template provided with each socket.

The Diamond Line of semiconductor hardware will also include a wide range of heat exchangers to control semiconductor operating temperatures and extruded

More Details? CHECK–OFF Page 120

aluminum heat sinks designed to deliver increased power dissipation per unit of cost.

The heat exchangers are available with universal hole patterns for various semiconductor types and TO-5 case configurations. A universal heat dissipator for all power tab plastic transistors, SCRs and triacs will be introduced to the line. Priced at 68 cents, the clip and dissipator assembly may be mounted vertically or horizontally as circuit density dictates.

Mounting kits for transistors and rectifiers will include hardware, insulating mica washers, wiring lugs and other components for application requiring electrical isolation.

A component carrier for mounting discrete components and integrated component packages rounds out the semiconductor hardware line and enables the hobbyist to create his own dual-in-line ICs. The carrier plugs into standard 0.100 x 0.300-inch dual-in-line sockets or into printed-circuit boards with the same grid pattern.

The component carrier is available with a snap-on cover. Clearance for components within the cover is keyed to accept a broad range of component and package sizes. Heavy duty pins accept wires and leads to size 24. Available at your local electronics dealer. For more information use *check-off* on page 120.

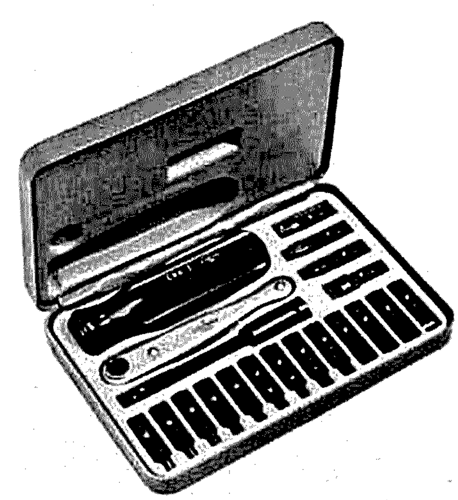
radio-electronics hobby projects

This new book by the editor of *Radio-Electronics* magazine provides a unique assortment of 32 practical electronics projects for the experimenter. For technicians the book includes projects on a fet dual-trace scope switch, a 3-way waveform generator, a scope calibrator, an audio tone-burst generator, a low-voltage electrolytic tester, a dot/bar generator and an fm stereo multiplex generator.

For the hi-fi buff there are multiplex tuners and adapters, stereo amplifiers and modifiers, mixers and speaker systems.

Musically-inclined experimenters will find vibrato and tremolo devices, rhythm lights plus a complete synthesizer. The auto enthusiast will find devices such as an electronic ignition system, a road icing alarm and an automatic windshield-wiper pause-controller. For the home handyman there are timers, a tape-slide synchronizer, a phototach, and an IC sound relay, among others. Each project includes a detailed parts list and a complete schematic or working drawing, plus explicit instructions on building, calibrating and operating the completed unit. 192 pages, many illustrations. $6.95 hardbound, $3.95 paperbound from TAB Books, Blue Ridge Summit, Pennsylvania 17214. For more information use *check-off* on page 120.

midget ratchet kits



The Chapman Manufacturing Company has a complete line of midget ratchet kits that should be especially interesting to the home electronics experimenter. These kits include all types of difficult-to-find drive tools including Bristol multiple-spline adapters, Allen head adapters, Phillips head adapters, Reed and Prince adapters and slotted-head adapters, as well as a ¼-inch drive adapter. The Allen head adapters are available in both English and metric sizes.

The Chapman adapters are constructed of high-strength chrome nickel molybdenum alloy steel and are precision made to insure a proper fit. Their dual-

purpose knurled "spinner tops" allow for quick finger tightening of threads and for instant "push-down" removal of the adapter from the ratchet. The Chapman ratchet, designed to function in confined areas, operates with minimum movement of the handle.

The 6320 tool kit shown in the photograph above includes 1 midget ratchet, 1 extension, 1 screw-driver handle, 12 Allen hex adapters, 2 slotted-head adapters, 2 Phillips head adapters and a ¼-inch drive adapter. The 6320 kit is priced at $12.95 at your local distributor. Manufactured by The Chapman Manufacturing Company, Route 17, At Saw Mill Road, Durham, Connecticut 06422. For more information use *check-off* on page 120.

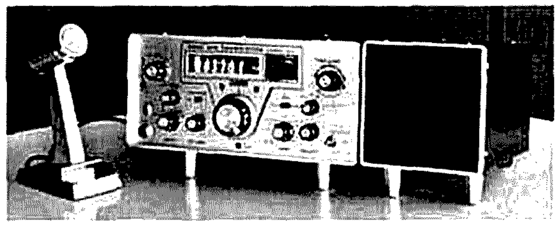
new alkaline batteries

A new generation of alkaline batteries, so good that they can outperform all previous types by as much as 20 to 25 per cent, has been introduced by the Mallory Battery Company. The new Duracell* batteries have higher energy capacity than any other alkaline batteries available up to now, made possible by an entirely different internal construction, with fewer parts allowing for an increased volume of energy-producing materials in the battery.

Featuring a new copper and black label design, the batteries will be manufactured in all popular sizes for use in radios, cameras, flashlights, tape recorders, and other consumer products. Available at your local dealer.

*Duracell is a registered trade mark of P. R. Mallory & Co., Inc.

ssb transceiver



The Robyn International Model 500 high-frequency ssb transceiver features 6-digit frequency readout with 100-Hz accuracy. The transceiver runs 500 watts PEP on all bands, 80 through 10 meters and boasts a complement of over 125 semiconductors and six vacuum tubes in addition to the six Nixies in the digital readout. An ac power supply, speaker and microphone are included in the price of $895; an optional 12-Vdc supply is available.

For more information write to Robyn International, Box 478, Rockford, Michigan 49341 or use *check-off* on page 120.

vhf scaler

The new Dycomm digital vhf frequency divider/prescaler, the model PSU-13, is a high-sensitivity unit with a divide-by-ten scaling factor. The unit operates over a minimum frequency range of 10 to 240 MHz. Inherently sensitive, it will operate properly throughout its frequency range with input levels under 500 mV, and is guaranteed to operate at 180 MHz with an input level of 100 mV.

Advanced circuitry and design are featured in the PSU-13. The heart of this circuitry is a custom medium-scale integration integrated-circuit chip. Other notable features include a high output level of 2 volts peak-to-peak (minimum) across an open circuit, with typical output levels of 3.5 volts. This feature is enhanced by a capability to drive up to 2 feet of coaxial output cable while measuring or dividing frequencies in excess of 150 MHz. The PSU-13 has proven extemely satisfactory when used as a prescaler for Monsanto, Heath, Hewlett-Packard and other counters. Those not owning counters may use the PSU-13 with a calibrated communications receiver to obtain relatively accurate fre-

More Details? CHECK—OFF Page 120

quency measurements. The PSU-13 will also serve to sync vhf signals with oscilloscopes having frequency responses in the 10- to 30-MHz range.

Sold complete and ready to operate (not a kit), the PSU-13 is factory-adjusted for maximum sensitivity over its entire range. All adjustments are internal and need not be continuously varied by the owner.

Complete with self-contained 110-Vac power supply the PSU-13 weighs less than 1½ pounds. Attractively finished in modern flat black with white lettering and power cord, the PSU-13 is a natural complement for any existing frequency counter or communications equipment. $89.95 from Dynamic Communications, Inc., Post Office Box 10116, Riviera Beach, Florida 33404. For more information use *check-off* on page 120.

solid-state battery

The Mallory solid-state battery was developed as a highly reliable, extremely long shelf-life power source. The anode is lithium and the cathode a metal salt; the electrolyte is a lithium ion-conductive electronically insulative *solid.* The electrolyte also serves as the separator between the anode and cathode. The reactive nature of the materials used in these new batteries requires that they be hermetically sealed. The absence of any liquid in the system completely eliminates corrosion or gassing.

Construction of the new batteries is simple: the cells are fabricated by pressing the individual components together. The cells are stacked in a suitable container and sealed with a hermetic cap. No containers are needed as in conventional battery; this feature is ideal for high-voltage low-current power sources.

Mallory solid-state batteries are currently available only in research quantities. Interested parties should contact Mallory Battery Company, Tarrytown, New York 10591 for information and discussion of their particular requirements.

audio power module

Sinclair Radionics announced the first integrated circuit amplifier and preamplifier, the IC-10, nearly three years ago. Integrated circuit production technology has improved tremendously during this period of time, and now Sinclair/Audionics is proud to introduce the new IC-12.

The IC-12 offers a number of important advantages over the IC-10 and other similar devices now available at low cost to the hobbyist or commercial user. Perhaps most important is the useful amount of power available combined with very low harmonic distortion. Of equal importance is the ease in which the IC-12 may be used. Each IC-12 is supplied with a comprehensive instruction manual and a predrilled and etched circuit board. The circuit board accepts the required external components and the IC-12. The IC-12 is supplied in a 16-pin dual-inline package and may be used as an integral part of other circuitry. In addition, no tricky initial setup bias adjustments are required; simply add the additional external components and the IC-12 is ready to perform as a high-gain wideband audio amplifier.

The IC-12 is basically an operational amplifier with a quasi-complementary output stage. Power supply requirements are thus simple and inexpensive. The idle current consumption of the IC-12 is only 8 mA making batteries practical as a power source. The Sinclair PZ-5 supply is also ideal for use with the IC-12 and will power a pair to rated output. Aside from having simple setup and power supply requirements, the IC-12 is also very rugged and stable. No addtional heatsinking is required; an extruded aluminum fin is part of the package and is more than adequate under all normal operating conditions.

Power output of the IC-12 is 6 watts into 8 ohms (with a 30-volt power supply). Total harmonic distortion is less than 1% at any audible audio frequency. Frequency response is 5 Hz to 50kHz ±1 dB, depending upon the values of the

More Details? CHECK—OFF Page 120

external components. Total device gain is 90 dB; noise is −70 dB or better. The IC-12 is priced at $8.95 from authorized dealers, or from Audionics, Inc., 8600 Northeast Sandy Boulevard, Portland, Oregon 97220. For more information use *check-off* on page 120.

electronics self-taught with experiments and projects

Rather than simply telling you how to build some electronic gadget, and leaving you wondering why he bothered, this unique beginner's guide to electronics by Jim Ashe offers a much more worthwhile challenge, and greater assurance of gaining useful knowledge from its use. Written especially for serious experimenters, hobbyists and students this book shows *why* certain things are done, tells *how* devices and circuits work, and suggests innovations that spur the reader forward to more important work in electronics.

Using what he calls a "new" electronics, author Ashe puts things in a different light than other writers — a fresh perspective that helps you see certain fundamentals clearly and gain new insights into *today's* electronics. The author tells how to set up a home lab (for around $20.00 for those who are thrifty) and what tools and equipment are necessary.

The book divulges many fascinating tricks with ordinary garden variety diodes and transistors. Expensive, special-purpose types are not needed. In fact, the author continually stresses the importance of keeping things simple and inexpensive. There are circuits and projects using integrated circuits, plus communications devices, logic circuits, industrial devices and lab-type equipment the reader can build and calibrate himself. $7.95 hardbound; $4.95 paperbound, from TAB Books, Blue Ridge Summit, Pennsylvania 17214. For more information use *check-off* page 120.

short circuits

versatile vox

In fig. 1 of this article, which appeared in the July 1971 issue, the transistors labeled 2N222A should be 2N2222A. Relay K2 should have been identified as a Hi-Vac type HC-1.

The receptacles on the output side of K2 should be labeled (from the top) "receiver," "antenna," and "P. A. tank." Also in fig. 1 the terminal labeled "to keyline" should be disconnected from the solder dot in the line between S1B and K1A and instead connected to the junction of R3 and the 0.01-μF capacitor.

high-power linear

In fig. 1, page 59 of the April, 1971 issue, the control-grid bias supply should be grounded at the positive end of the 40-μF filter capacitor nearest the 25k potentiometer.

deluxe mosfet converters

In fig. 1, page 42 and 43 of the February, 1971 issue there should be 10-pF capacitor between the rf amplifier on page 42 and the mixer on page 43. Piston trimmers are JFD VAM010W or Johanson 2951.

RTTY multimeter

In the circuit for the RTTY multimeter in fig. 13 on page 29 of the March, 1971 issue, the 20k resistor should be 200k. Also, the duty cycles quoted are incorrect; these should be approximately 33% for Teletype and 29% for Western Union. Calibration remains correct for standard Teletype machines.

sonobaby

Two of the diodes in the Sonobaby fm transmitter in fig. 2 on page 12 of October, 1971 issue are reversed — the two diodes on the right-hand side of the full-wave bridge should be turned around for proper operation.

ham radio

index to volume IV – 1971

This index covers all articles published in *ham radio* during 1971. The articles are listed alphabetically under each category along with the author, page number and month. Categories are: antennas and transmission lines; commercial equipment; construction techniques; fm and repeaters; integrated circuits; keying and control; measurements and test equipment; miscellaneous technical; power supplies; receivers and converters; RTTY; semiconductors; single sideband; transmitters and power amplifiers; vhf and microwave. Articles followed by (HN) appeared in the *ham notebook.*

antennas and transmission lines

audio

commercial equipment

construction techniques

fm and repeaters

integrated circuits

keying and control

measurements and test equipment

miscellaneous technical

power supplies

receivers and converters

RTTY

semiconductors

single sideband

transmitters and power amplifiers

vhf and microwave